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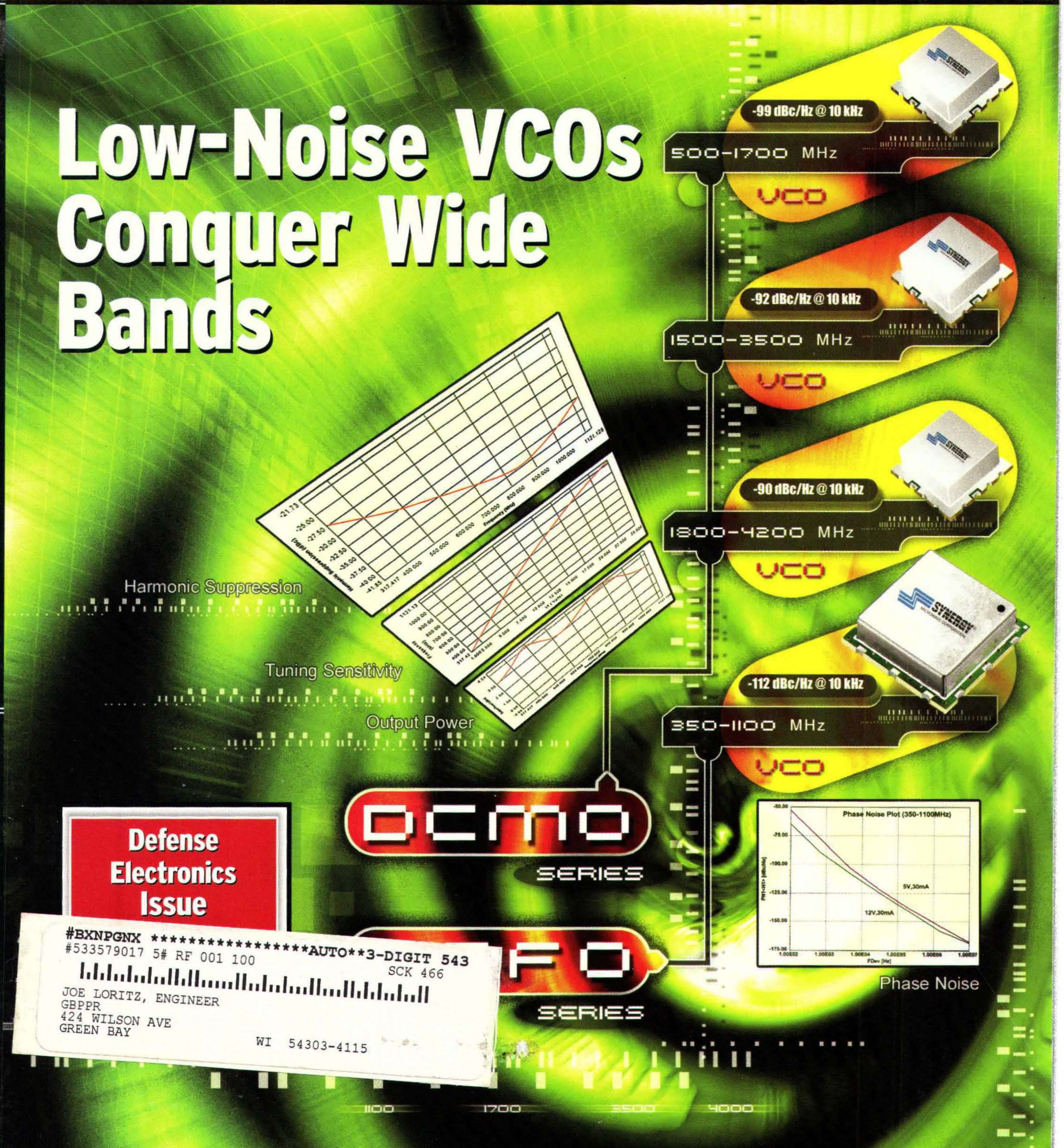
News
Narda Microwave celebrates 50 years

Design Feature
Antenna snares GPS/WLAN signals

Product Technology
Fast synthesizers switch 5 MHz to 20.48 MHz

Low-Noise VCOs Conquer Wide Bands



Defense Electronics Issue

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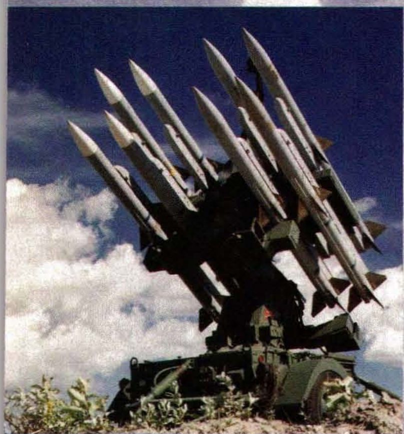

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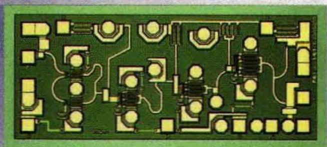
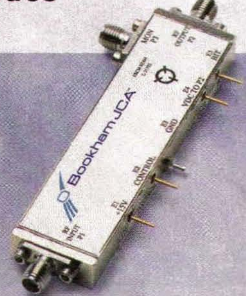
MMICs NEW

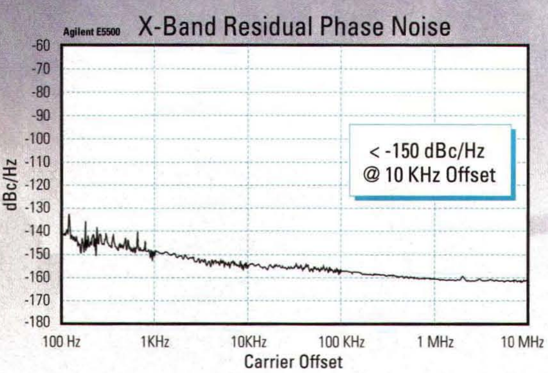
Model	Freq. Range GHz	Gain dB	N/F dB	P1dB dBm	V V	I mA
P35-5104-000-301	2–20	10	4.0	13	3.5	70
P35-5114-000-200	20–32	21	2.2	7	2	48
P35-5122-000-200	8.5–10.5	18	–	25	5	270
P35-5123-000-200	20–26	12	–	23	4.5	140
P35-5127-000-200	25–30	10	–	22	4	140
P35-5140-000-200	20–40	20	–	21	4.5	192





Broadband Power Amplifiers

Model	Freq. Range GHz	Gain dB min	N/F dB max	Flatness +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ
JCA218-3000	2.0–18.0	23	4.0	2.5	23	28
JCA218-3001	2.0–18.0	30	4.0	2.5	25	30
JCA218-3002	2.0–18.0	34	4.0	2.5	27	32
JCA218-4000	2.0–18.0	33	4.0	2.5	23	28
JCA218-4001	2.0–18.0	40	4.0	2.5	25	30
JCA218-4002	2.0–18.0	44	4.0	2.5	27	32
JCA218-5000	2.0–18.0	43	4.0	2.5	23	28
JCA218-5001	2.0–18.0	50	4.0	2.5	25	30
JCA218-5002	2.0–18.0	54	4.0	2.5	27	32
JCA618-4001	6.0–18.0	40	5.0	2.0	33	40



Low Noise Amplifiers

Model	Freq. Range GHz	Gain dB min	N/F dB max	Flatness +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ
JCA12-1000	1.2–1.6	25	0.8	0.5	10	20
JCA12-3001	1.0–2.0	40	0.8	1.0	10	20
JCA14-400	1.0–4.0	40	0.9	1.5	15	25
JCA34-301	3.7–4.2	30	1.0	0.5	10	20
JCA48-4001	4.0–8.0	42	1.0	1.5	15	25
JCA910-3000	9.0–9.5	25	1.3	0.5	13	23
JCA812-5001	8.0–12.0	45	1.5	1.5	10	20
JCA1218-5001	12.0–18.0	48	1.7	1.5	10	20
JCA1819-3001	18.1–18.6	25	2.0	0.5	10	20
JCA2021-3001	20.2–21.2	25	2.5	0.5	10	20

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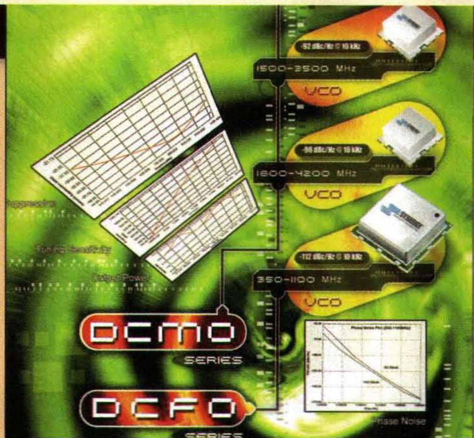


COVER STORY

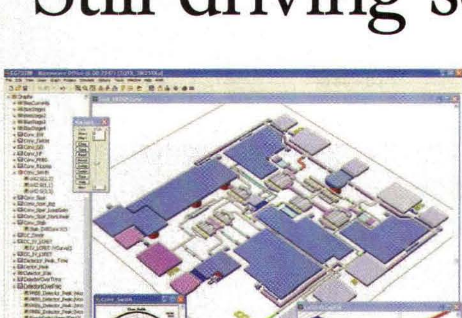
News

Design

Product Technology

AWR

Still driving software from another era?



If you're just trotting along with your designs, maybe you're using the wrong vehicle. Truth is, it's difficult to win the race-to-market riding an obsolete chassis. And sky-high maintenance fees aren't likely to deliver any magic upgrades to the 21st century. So trade in your old horsepower for the new standard in high-frequency design: Microwave Office™ 2003. It's the only solution that combines a modern unified database with the industry's most intuitive interface. And it's typically 2–10× faster than more antiquated solutions. See how efficient RF and microwave engineering can be. Why not take a test drive? Download a 30-day demo from www.mwoffice.com or call us at 310-726-3000 for more details.

Microwave Office 2003

MRF PQ Correction

▸▸ON P. 12 OF THE April 2004 issue of *MRF Product Quarterly* (Components section), an American Microwave Corp. photo mistakenly ran in an American Microwave, Inc. product module. We apologize to American Microwave Corp., Anaren Microwave, Inc., and to the readers of *MRF Product Quarterly* for the error.

The Editors of MRF Product Quarterly

Pardon My French, But . . .

▸▸WHILE ATTENDING the HYPER 2004 in France, I heard something disturbing from more than one credible source. Both Thales and Alcatel are actively discouraging and, in some cases, not allowing US parts to be designed into their military systems.

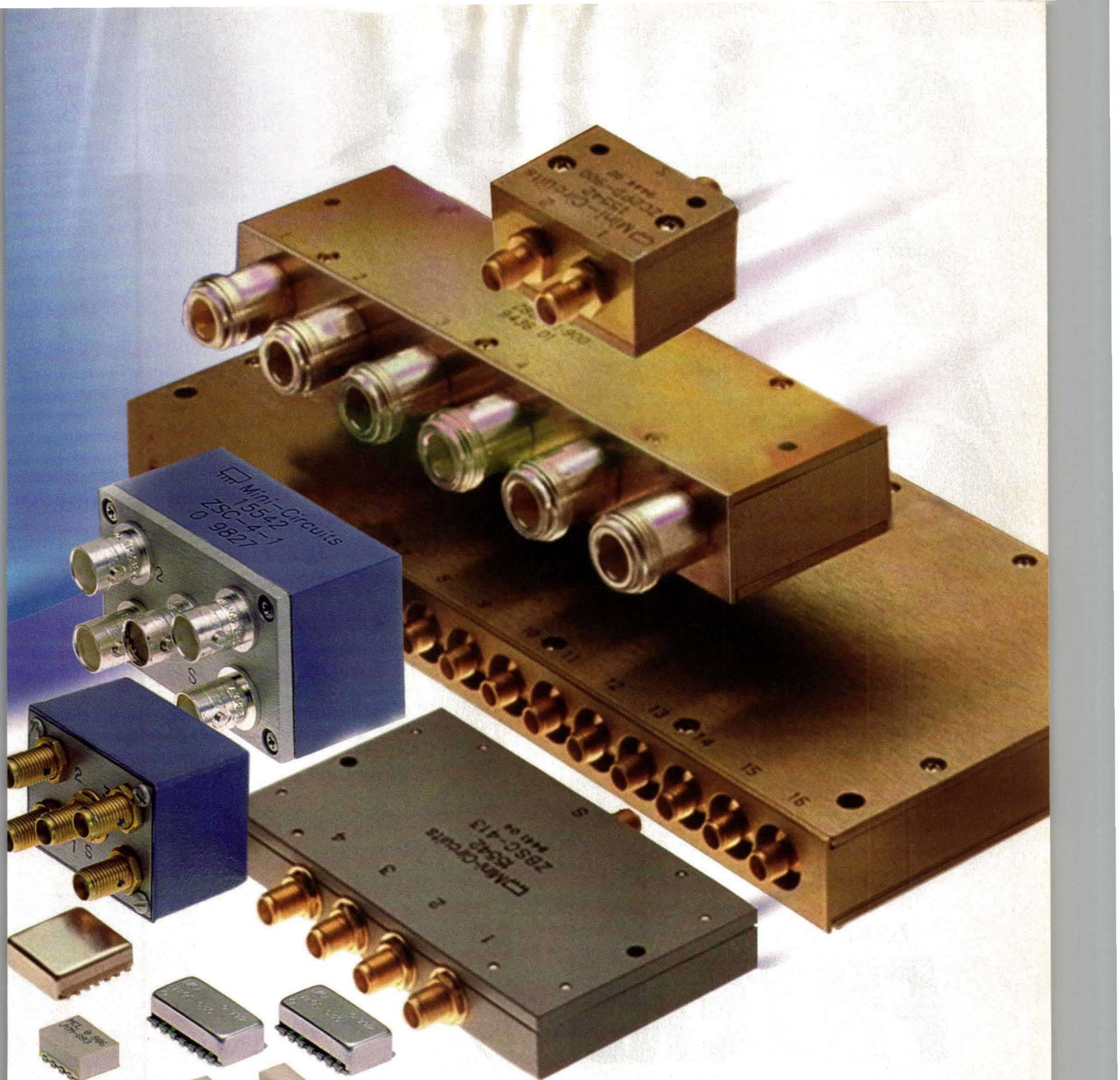
Considering that US export rules for RF and microwave components can carry end use and end user information, these companies have decided that they will not tell the US about the end use or give end user information about their military systems. Why do you think that they have this position? There was even a case in which a French component manufacturer had used a US integrated circuit as part of the total design, and when the French system house opened the unit and saw this, they strongly requested that it be designed out.

The position of these companies at present is to buy French first, European second, and Asian or Australian third unless they have no choice but to buy American. During the 1970s and 1980s, the French microwave community were buying a considerable amount of US products. The American companies were being very aggressive in marketing to them and offered, in many cases, no non-recurring engineering (NRE) charges in efforts to get designed into a system. This strategy worked, and it bore out that some US manufacturers had as much as 10 percent, if not more, of their sales coming from France. This was not without consequences for the French manufacturers. Many small French RF and microwave companies either went out of business or were bought out by larger firms.

Some major French companies want to focus on buying domestic, but they have, in the past, destroyed their own infrastructure. American companies have seen their business in France decrease, and the animosity is building.

The reality is that the US has a short memory when it comes to international affairs. We have put the "French" back into French fries, the local liquor stores are again selling Grey Goose vodka,

(continued on page 38)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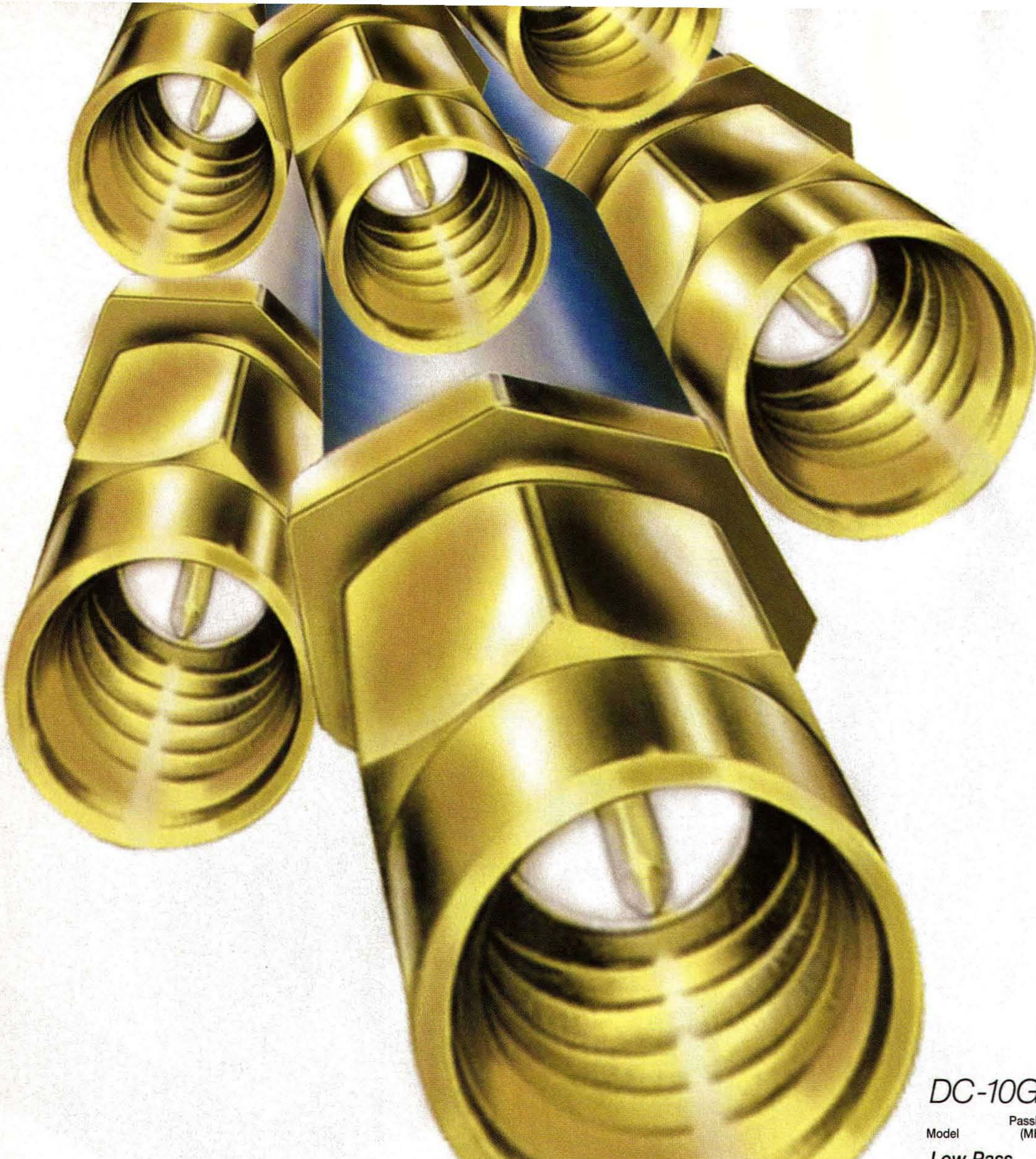
RF/IF
Designer's
Guide



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VLF-400	DC-400	560	660	20.95
VLF-490	DC-490	650	780	20.95
VLF-530	DC-530	700	820	20.95
VLF-575	DC-575	770	900	20.95
VLF-630	DC-630	830	970	20.95
VLF-800	DC-800	1060	1225	19.95
VLF-1000	DC-1000	1300	1550	19.95
VLF-1200	DC-1200	1530	1800	19.95
VLF-1700	DC-1700	2050	2375	19.95
VLF-2250	DC-2250	2575	2850	19.95
VLF-5000	DC-5000	5580	6600	19.95
VLF-6000	DC-6000	6800	8300	19.95
VLF-6700	DC-6700	7600	8900	19.95
High Pass				
VHF-650	850-2490	650	480	19.95
VHF-740	900-2800	740	550	19.95
VHF-880	1060-3200	880	640	19.95
VHF-1200	1340-4600	1180	940	19.95
VHF-1300	1510-5000	1300	930	19.95
VHF-1320	1700-5000	1320	1060	19.95
VHF-1500	1700-6300	1530	1280	19.95
VHF-1600	1950-5000	1600	1290	19.95
VHF-1760	2100-5500	1760	1230	19.95
VHF-1810	2250-4750	1810	1480	19.95
VHF-1910	2200-5200	1910	1400	19.95
VHF-2000	2410-6250	2000	1530	19.95
VHF-2100	2500-6000	2100	1530	19.95
VHF-2275	2640-7000	2275	1770	19.95
VHF-2700	3000-6500	2500	1800	19.95

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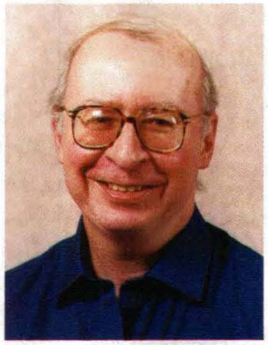
See our 244 page RF/IF Designer's Guide in EEM (Electronic Engineers Master)

MEMS Make Noise At MTT-S



"EXPERIMENTAL" TECHNOLOGIES ARE often slow to gain ground among high-frequency engineers. For example, high-electron-mobility-transistor (HEMT) devices, which are commonplace today, were used by only the most courageous of engineers during the early years of the technology. At present, microelectromechanical systems (MEMS) devices are viewed as a technology with high associated risk, although evidence from the recent Microwave Theory & Techniques Symposium (MTT-S) in Fort Worth, TX hints that greater acceptance of MEMS may be forthcoming.

While questions may remain among engineers considering MEMS devices for their designs, reliability does not appear to be one of them.

MEMS devices can be thought of as silicon integrated circuits (ICs) with moving parts. The technology has existed since the 1970s in the form of sensors, but RF devices have been rare. The most common RF device is the MEMS switch (which sacrifices the speed of a PIN diode for greatly enhanced isolation). MEMS microwave switch technology has existed since the 1980s, developed by Dr. Larry Larson of Hughes Research Labs (Malibu, CA) with support from DARPA.

Any new technology must provide advantages over an existing technology before it can replace the older approach. While MEMS switches, for example, offer those advantages over other technologies in terms of size, power handling, power consumption, and their ability for high levels of integration on silicon, they pose a mystery for many RF/microwave engineers in terms of reliability. According to Dan Hyman, President of XCom Wireless (www.xcomwireless.com, Signal Hill, CA), who spoke recently at a MEMS-inspired panel session at the MTT-S, his company's work on MEMS switches, switch matrices, filters, phase shifters, and antenna assemblies has shown the technology to be extremely reliable. He and his designers are such firm believers in the technology, in fact, that he noted a recent development project for a front-end design in which the majority of the passive signal-routing components were MEMS devices.

On the MTT-S exhibit floor, Dow-Key Microwave (www.dowkey.com, Ventura, CA) showed results for accelerated lifetime testing of MEMS switches, with almost imperceptible degradation in electrical performance even after millions of switching operations. The company also introduced their model M1C06-CDK2 single-pole, double-throw (SPDT) DC-to-6-GHz switch rated for an amazing 100 million cycles.

While questions may remain among engineers considering MEMS devices for their designs, reliability does not appear to be one of them. The technology is not the answer for all applications (the lack of speed being an essential limiting factor), but it is an approach worth considering.

Jack Browne

Publisher/Editor

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A Penton Publication

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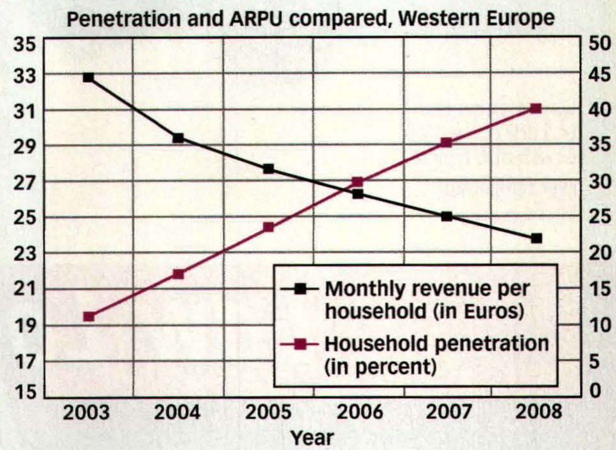
News items from the communications arena.

European Broadband Penetration Set To Soar To 40 Percent By 2008

BOSTON, MA—According to *EMEA Consumer Fixed-Line & Media Forecast*, a report from the Yankee Group, 2003 marked a watershed year for broadband services in Europe. Four important developments transformed the market:

- Most operators introduced entry-level tariffs to maintain or grow their customer base and market share. Some also introduced pay-as-you-go tariffs and other packages designed to lower entry barriers to broadband.
- A massive continent-wide marketing blitz firmly established the broadband message of speed and convenience in the minds of Internet users. In the UK, for example, 90 percent of consumers have heard of broadband.
- Many service providers made the first serious efforts to establish new revenue sources for broadband, resulting in a widening range of services (including bundled triple-play services) and making it easier to justify lower prices for the basic connection.



Source: The Yankee Group EMEA Consumer Fixed-Line & Media Forecast, First Quarter 2004

- Competition has heated up in most countries, driven by the perception that broadband is now the main event in consumer telecommunications.

Because of these highly positive market developments, the Yankee Group has revised upward its forecast for broadband penetration. They now expect broadband to connect to 40 percent of households in Western Europe by 2008 **(see figure)**. However, recent tariff charges will drag down ARPU, so revenue will not be substantially higher.

RFDomus Announces Introduction Of Q-MAX Technology

IRVINE, CA—RFDomus, Inc., a fabless semiconductor company that specializes in advanced RF and analog mixed-signal (AMS) semiconductor devices for the wireless-communications marketplace, has announced the introduction of its patent-pending Q-MAX™ technology.

Designed to bridge the gap between competing demands for low power consumption and high-performance, RFDomus' Q-MAX technology is based on a local-oscillator (LO) architecture that enables an order-of-magnitude reduction in power consumption when compared to conventional LO implementations. Designed for simplifying the integration of the LO block with other radio blocks in advanced complementary-metal-oxide-semiconductor (CMOS) and silicon-germanium (SiGe) processes, Q-MAX technology provides the flexibility needed for inclusion in complete single-chip RF transceivers for multiple wireless applications, including GPS, cellular phones, Bluetooth™, wireless LAN (WLAN), and ultrawideband (UWB).

"RFDomus' patent-pending Q-MAX technology is portable to multiple wireless-communications applications such as GPS, Bluetooth, wireless LAN, and ultrawideband," says Kevin Strong, RFDomus' executive vice president for business development. "Using the Q-MAX technology, we have designed an LO block for GPS applications that consumes ten times less power than an LO in competitive products. We are now in the process of designing an LO for Bluetooth applications, and we expect to achieve similar results."

The X Initiative Welcomes TSMC As Its Newest Member

MOUNTAIN VIEW, CA—The X Initiative has welcomed Taiwan Semiconductor Manufacturing Co. Ltd. (TSMC)—the world's largest semiconductor foundry—as the latest member of the semiconductor design-chain consortium. Having successfully verified the 0.13-µm X Architecture design rules with test chips, TSMC is now working with select customers on their circuits to leverage the performance, cost, and power advantages of the X Architecture. With the world-leading foundry leveraging its manufacturing facilities to fabricate X Architecture silicon, the innovative chip architecture is one step closer toward broad commercial adoption by the global semiconductor industry.

"TSMC is constantly evaluating new technologies that address nanometer design challenges and add value to our customers," comments Genda Hu, vice president of marketing at TSMC. "The promise of the X Architecture has prompted us to verify our design rules for the X Architecture at the 0.13-µm and below process nodes. Now we're engaging with select customers on their circuits that employ this new design-implementation approach."

The X Architecture represents a new way of orienting a chip's microscopic interconnecting wires using diagonal pathways, as well as the traditional right-angle, or "Manhattan," configuration. By enabling designs with significantly less wire length and fewer vias (the connectors between wiring layers), the X Architecture can provide significant improvements in chip performance, power consumption, and cost.

The pre-production phase of the design-to-silicon roadmap for the X Architecture, laid out by the X Initiative in 2002, was completed with the announcement of functional silicon results by an X Initiative member late last year. The focus of the X Initiative's collaborative design-chain preparation is now to enable broad adoption of the X Architecture for production manufacturing at current (130 nm, 90 nm) nodes and to demonstrate manufacturing scalability into future process nodes. First production chips are expected this year.

"We are very pleased to add TSMC to the X Initiative's roster," states Aki Fujimara, X Initiative steering group member and CTO for new business incubation at Cadence Design Systems, Inc. "Fabless semiconductor companies stand to reap significant improvements in design speed, power, and cost by leveraging the X Architecture. We're looking forward to working closely with TSMC and their customers to enable a key manufacturing path for X Architecture designs."

> *"Fabless semiconductor companies stand to reap significant improvements in design speed, power, and cost by leveraging the X Architecture."*

The First WiMAX Certified Broadband Wireless System Is On The Drawing Board

CALGARY, ALBERTA, CANADA AND SUNNYVALE, CA—Fujitsu Microelectronics America, Inc., a global ASIC and semiconductor-solutions provider, and Wi-LAN, Inc., a global provider of broadband wireless-communications products and technologies, have announced their joint goal to produce the world's first WiMAX Certified broadband wireless system. System-on-a-chip (SoC) engineering samples are planned for the fall of 2004, and the complete system is expected to be available for WiMAX Forum conformance and interoperability testing in the first half of 2005.

"Fujitsu Microelectronics' expertise in designing high-performance, highly integrated system-on-chip solutions that incorporate embedded processors and mixed-signal technology will be instrumental in bringing WiMAX Certified broadband wireless products to economical mass deployment," says Keith Horn, FMA's vice president of marketing. "Together, Wi-LAN and FMA have all the skills required to develop broadband wireless SoC products that will lead the market in WiMAX Certified systems."

"The timeline that Wi-LAN and FMA have set for first-to-market WiMAX Certified systems dovetails nicely with the WiMAX Forum's timeline for certification and interoperability testing," comments Dr. Sayed-Amr El-Hamamsy, Wi-LAN's president and CEO. "In spite of the hype about early equipment availability from other vendors, the fact is that there can be no certified products before WiMAX conformance specifications are established. The chairman of the WiMAX Forum 2-11GHz technical working group expects to complete conformance specifications in November 2004."

The WiMAX Forum is a non-profit corporation formed to help promote and certify the compatibility and interoperability of BWA (Broadband Wireless Access) equipment.

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S3W2	S3W5	N3W5	3	±0.40
S4W2	S4W5	N4W5	4	±0.40
S5W2	S5W5	N5W5	5	±0.40
S6W2	S6W5	N6W5	6	±0.40
S7W2	S7W5	N7W5	7	±0.60
S8W2	S8W5	N8W5	8	±0.60
S9W2	S9W5	N9W5	9	±0.60
S10W2	S10W5	N10W5	10	±0.60
S12W2	S12W5	N12W5	12	±0.60
S15W2	S15W5	N15W5	15	±0.60
S20W2	S20W5	N20W5	20	±0.60
S30W2	S30W5	N30W5	30	±0.85
S40W2	S40W5	N40W5	40	±0.85

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Hittite Adds New Sales Reps In China And Norway

CHELMSFORD, MA—Hittite Microwave Corp., a supplier of complete MMIC-based solutions for communication markets, announced the appointment of a new sales-representative firm for select customer accounts in the People's Republic of China and Hong Kong. Secom Telecom Co., established in June 1993, is headquartered in Shenzen, China and is one of the most experienced electronic-component distributors in China and Hong Kong. Secom's areas of expertise include: semiconductors and electronic equipment for data and voice communication, medical, and financial applications.

Together with Hittite's current representative, Wai Tat Electronics, and Hittite's Shanghai and Beijing offices, Secom will support Hittite's direct sales channels in China and Hong Kong by focusing promotion on a select group of key customers. Secom can be contacted via phone at (86-755) 25155888, via fax at (86-755) 25155816, or e-mail at sales@secomtel.com.

Hittite has also announced the appointment of a new sales representative firm to serve customers in Norway. Bredengen AS, headquartered in Oslo, Norway, is a full-service communication components company. Transmission and communication technology-related products have been a core business since Bredengen's parent company's inception in 1926. Bredengen's areas of expertise include: products for RF and microwave signal transmission and processing.

Bredenben can be reached by phone at +47 21 00 91 00, via fax at +47 21 00 91 01, or e-mail at bredengen@bredengen.no_.

New Company Offers 50 Years Of Experience With RF Connectors

MERRIMACK, NH—Coaxant, Inc. is a recently formed manufacturer of custom RF and microwave connectors, with particular focus on replacing other manufacturers' obsolete parts.

The Coaxant, Inc. team boasts over 50 years of experience in the manufacture of stainless steel, brass, and other specialty metal designs for markets including wireless, commercial communications, and military defense. The company offers a full range of cost-effective engineering capabilities, enabling it to meet customers' performance requirements to 18 GHz.

To learn more about Coaxant, Inc., please visit www.coaxant.com or e-mail: sales@coaxant.com.

Kudos

WARREN, NJ—ANADIGICS, Inc., a supplier of wireless and broadband solutions, announced that its co-founder, executive vice president, and CTO, Dr. Charles Huang, was honored at his induction into the High-Tech Hall of Fame for his pioneering contributions to the development of GaAs ICs. He is one of three people in the Outstanding Researchers category inducted into New Jersey's High-Tech Hall of Fame for 2004.

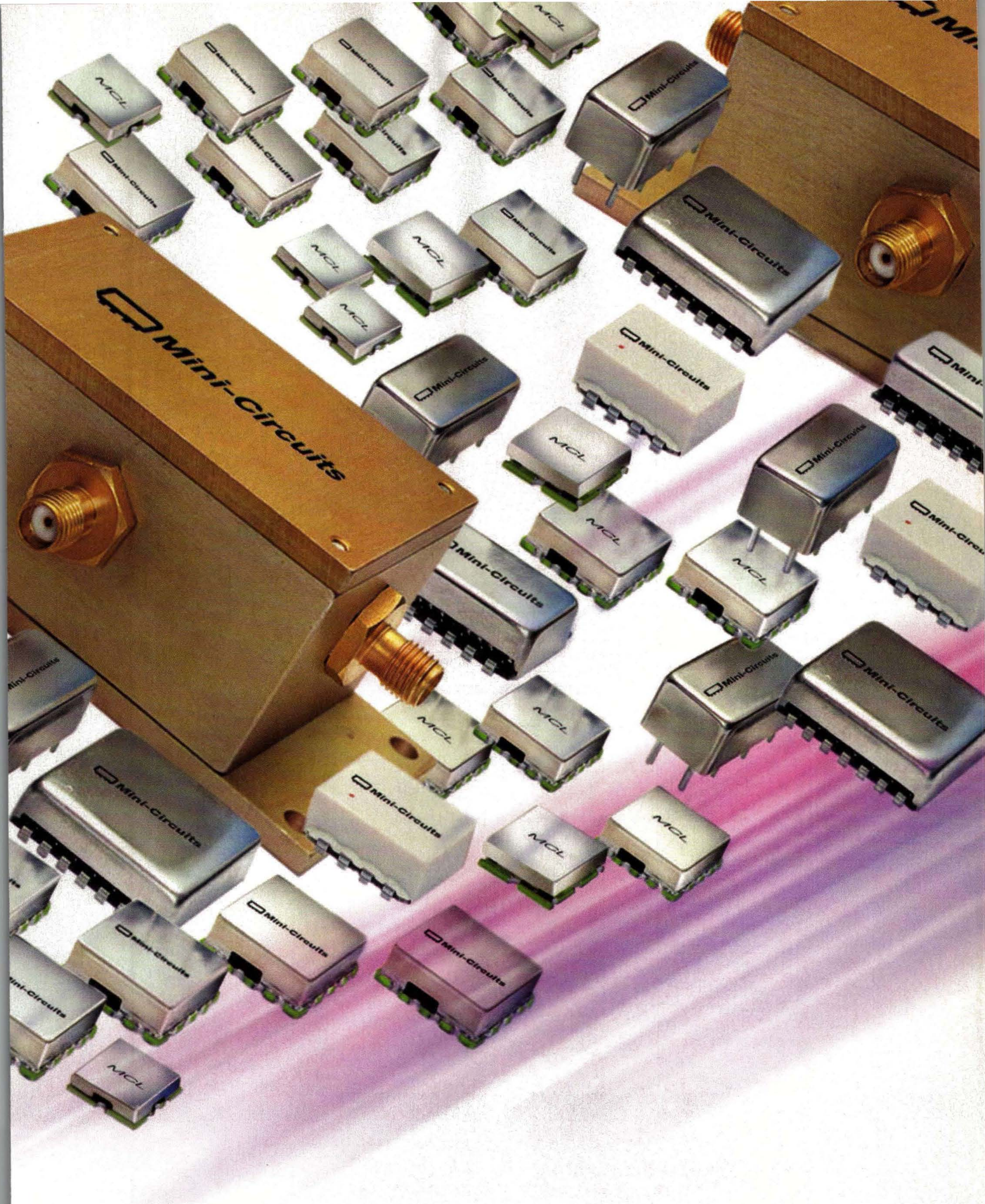
Dr. Charles Huang was honored at his induction into the High-Tech Hall of Fame for his pioneering contributions to the development of GaAs ICs.

Dr. Huang co-founded ANADIGICS and has served as its executive vice president and CTO since 1985. He has over 20 years of experience in the design, development, and production of components made using gallium-arsenide technology for companies such as Hewlett-Packard and Avantek. Before co-founding ANADIGICS, Dr. Huang was director of gallium arsenide research and development and wafer fabrication at Avantek.

ENDICOTT, NY—Endicott Interconnect Technologies (EI), a global supplier of products and solutions for the electronics industry has announced the successful completion of their annual ISO 9001:2000 surveillance audit. The annual review, conducted by Bureau Veritas Quality International (BVQI), was completed over a four-day period and touched all aspects of their quality system, including manufacturing systems, customer relationship management, and new product development.

WESTLAKE VILLAGE, CA—Inphi Corp., a manufacturer of analog mixed-signal electronic components for computing and communications, announced that it has received certification as an International Organization for Standardization (ISO) 9001:2000 Fabless IC company. Inphi received the certification for both engineering and manufacturing after an exhaustive review of its business practices and procedures.

Certification requirements include management leadership, a pro-active and well-trained work force, customer feedback, measurement, documentation, internal audits, continuous improvement, and third-party validation. MRF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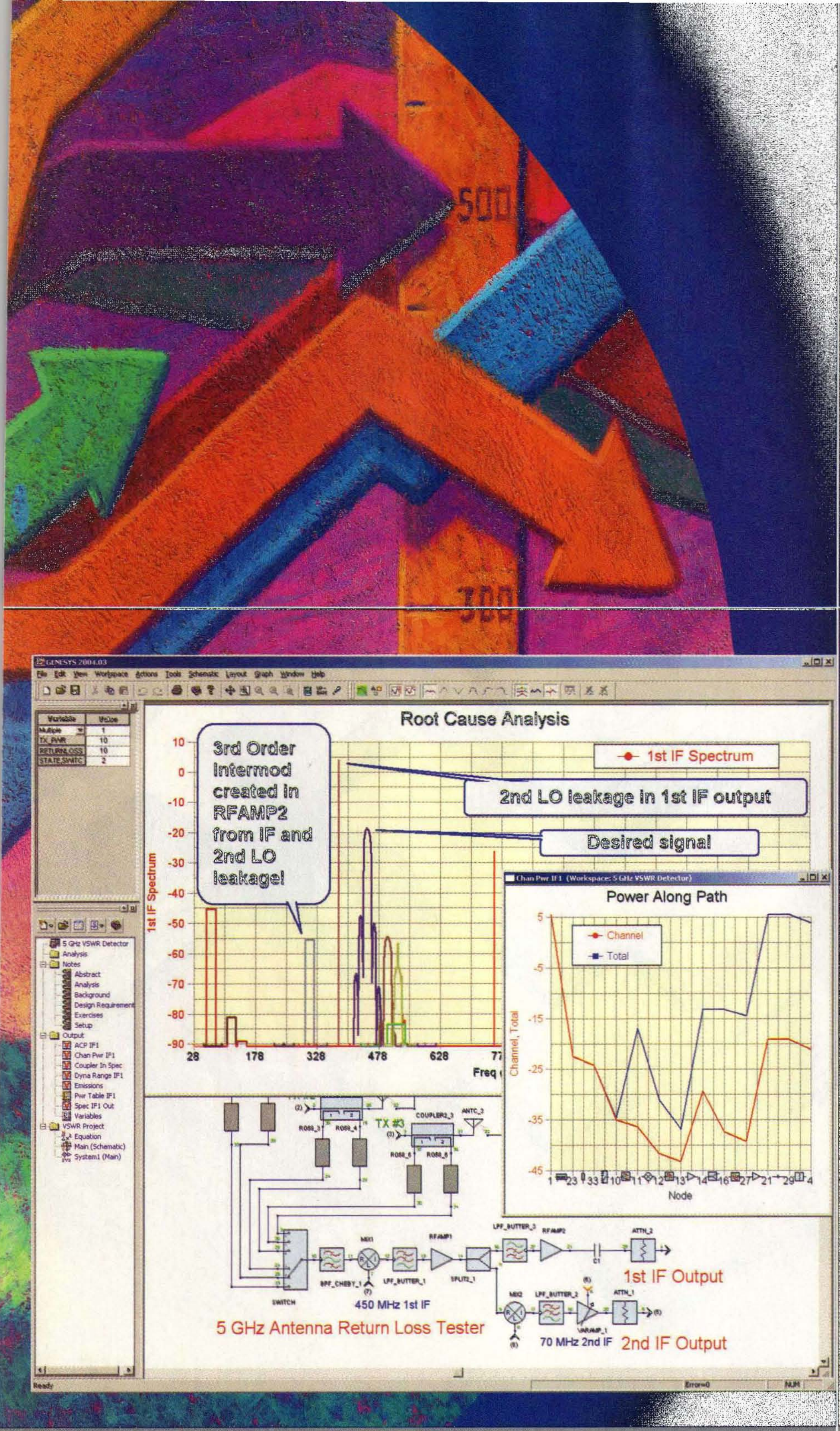

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70 MHz 2nd IF
2nd IF Output

NEWS

Narda Microwave Celebrates 50 Years

One of the high-frequency industry's longest-running success stories has diversified over its 50 years to become a leading supplier of microwave components and test instruments.

JACK BROWNE
Publisher/Editor

Fifty years ago, Narda Microwave became a corporation. What began as a self-funded experiment in starting a business between three engineers and a businessman has transformed over time into one of the most trusted names in the microwave industry. Very much representative of the industry itself, the company has shown the resilience and flexibility over the years to weather the harsh, cyclical nature of the high-frequency-electronics business. Narda has been a strong supplier to both military and commercial customers for its high-performance components and assemblies, and is nearly synonymous with industrial radiation-monitoring equipment and personal electromagnetic (EM) safety monitoring with its Nardalert XT and RadMan XT portable personal monitoring devices.

The company actually began in 1953, started by three engineers, Bill Bourke, Jim McFarland, Stu Casper, and John McGregor, a writer and businessman. The four had earned a $20,000 contract with Kollsman Instrument to supply engineering services at Kollsman. While not at Kollsman, Stu Casper and Jim McFarland were designing instrumentation for X-band use while Stu also devised a frequency meter for wideband measurements. Bill Bourke and Jim McFarland were also designing a variety of passive components for high-frequency use, including attenuators and terminations. Working in a small rented space in Mineola, NY in the back of a machine shop, they subcontracted prototypes to the machine shop and ran the company on funds withdrawn from their own bank accounts. Early products (and sales) included an X-band slide screw tuner and various attenuators, fabricated by machine shops according to their engineering drawings.

This was one of four buildings housing Narda Microwave in Mineola, NY.

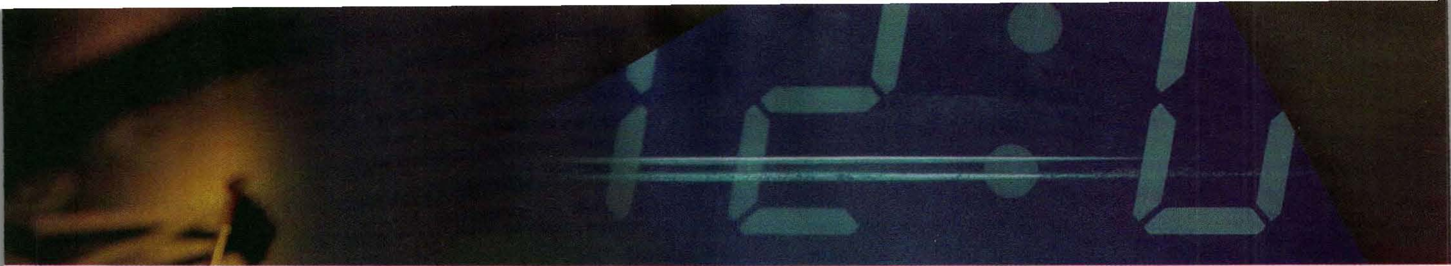
The first "high-level" product was the model 802 frequency meter, manufactured under contract by Peerless Instruments. In October 1953, the fledgling company sold one high-power impedance meter to Maxson Corp. and another five to Bell Telephone. At a unit price of $1700, these sales gave

the company a then-record sales month of $8500 in bookings. As the funds from the Kollsman Instrument contract ran out, the partners found themselves working without salaries toward the end of 1953. Because they required additional test instruments to grow their product lines, they secured personal loans from banks totaling $12,000 to pay for the needed test gear.

The new test equipment enabled the partners to develop a line of bolometers that were announced in January 1954. Two months later, they would hire their first employee, and on July 1, 1954, they incorporated their company as The Narda Corp. Three years later, the name would be changed to The Narda Microwave Corp. By December of 1954, sales exceeded $10,000 per month. By late spring of 1955, they had outgrown their rented 1600 sq. ft. of space and moved to a larger site (about 6000 sq. ft.) at 160 Herricks Rd. in Mineola, NY. By the end of June that year, sales for the year had reached $116,000, growing to an annual total of $321,000 for 1956, and tripling to $891,000 by end of June 1957. By 1957, the company had 75 employees, including 11 engineers. That year, the partners made their first acquisition, a company called Kama Instrument, bringing the total number of products for the company to 325.

On August, 16, 1957, Narda made its first public offering of company shares, selling 90,000 shares at $3 per share. (Prior to that, the partners had offered shares in the company, but only to employees.) With the help of the finances from the public offering, Narda Ultra Sonics Corp. was formed on October 1, 1957 to design and market industrial ultrasonic cleaners. The following September, The Narda Hydraulics Corp. was formed to develop electronic equipment for government and commercial customers. By 1959, the company was housed in four buildings in Mineola **(see figure)** totaling 20,000 sq. ft. It offered 700 products and employed 125 people.

By 1960, annual sales exceeded $2,250,000, and the company had launched numerous successful products, including a high-power microwave modulator, a ferrite isolator, precision attenuators, and a digital frequency meter. The following year, annual sales had reached $2,651,435 and construction was completed on a 40,000-sq.-ft. facility in Plainview, NY. The company catalog now required 170 pages. In 1965, Narda acquired the Microline product line from the Sperry Rand Corp. The following year,



annual sales exceeded $4,474,000. By 1967, the company's application for a listing on the American Stock Exchange is approved, and ground is broken on a 40-acre plot purchased in Melville, NY. New products that year included the model 4016 miniature Ku-band coupler, the model 3074 precision reflectometer coupler, and the model 445 precision RF power bridge.

By 1968, the new building was completed and occupied, and computers were now being used in the form of newly installed automated milling and drilling machines. The following year, Narda formed a subsidiary called Narcom (headed by Stu Casper) to sell commercial- and military-communications systems. That March, at an IEEE show, the company introduced the world's first commercial solid-state swept frequency generator. In addition, the company's new model 8100 Surveyor was acknowledged as one of the country's top 100 industrial products that year. Sales for fiscal 1969 reached $6,110,000.

In 1971, Bill Bourke was elected the company's Chairman of the Board. The following year, Nardacom was discontinued. In 1973, Narda France was formed to serve growing European markets. By 1974, export sales would exceed $2 million, and Cayuga Associates would become a subsidiary. The following year, after introducing lines of Gunn oscillators and PIN diode switches, annual sales exceeded the $10 million mark. Also in 1975, Narda purchased certain assets of a company called Anacom, Inc. (Santa Clara, CA), providing the basis for Narda's Pacific Operations to begin business that December is a new 6300-sq.-ft. facility in Santa Clara, CA. By 1977, the Pacific Operations would move to a larger facility in Sunnyvale, CA. In 1979, the company's 25th year, the firm introduced the world's first 6-to-18-GHz low-noise amplifier (LNA), and sales grew beyond $17 million.

By 1980, annual sales exceeded $20.7 million. That October, ground-

breaking ceremonies took place for a new facility to be built on an 18-acre plot in Hauppauge, NY. In 1982, Narda's Pacific Operations, now known as Narda West, moved to a new 45,000 sq.-ft. facility. For the company as a whole, sales for the year exceeded $27,912,000.

In 1983, Narda was purchased by Loral Corp. for about $45 million. Narda continued to develop high-performance safety products and components for military system, but also expanded its product lines to meet the needs of a growing wireless-communications market, notably with the 72000 series of VSWR monitors, also known as the "CATS" Communications Antenna Test System.

In 1994, even though Lockheed Martin acquired Loral Corp., the Narda name survived and product sales continued strong. In 1998, a group of former Loral executives purchased 10 divisions of the former Loral Corp.—including Narda—and formed L-3 Communications.

In 1999, L-3 purchased the former Wandel & Goltermann Safety Test Solutions operation in Germany from Wavetek, combining those products with Narda's safety products under the Narda Safety Test Solutions name. In 2001, Narda introduced the Nardalert personal monitor. In 2003 came the SRM-3000 selective radiation meter. By November of that year, the Satellite Networks division of L-3 was merged into the Narda Microwave-East operation in Hauppauge, NY.

Today, Narda continues to develop innovative products in a wide variety of technology and market areas, such as the model 4229-10, a single broadband 10-dB coupler capable of operating from 1 to 40 GHz; the model PCSW11799-12, a DC-to-8-GHz electromechanical switch that fits in a package measuring only 0.79 × 0.75 × 0.4 in. and the SRM selective radiation meter, a powerful handheld instrument capable evaluating EM emissions on multiple sources from100 kHz to 3 GHz. MRF

(continued from page 13)
and we will always have the lovely "Lady of Liberty" in New York Harbor. I think that we should get over it, and so should the French. We need to put our political differences aside and go back to building products, selling them, buying them, and moving the big ball forward.

Name Withheld By Request

Editor's Note: Thanks for your comments. Would any of the readers like to comment on what's been written in the letter above? Please e-mail us at jbrowne@penton.com.

Academic Papers

I ENJOYED READING Jack Browne's Editor's Note in the April 2004 issue of *Microwaves & RF* ("Making A Better Microwave Show," p. 17). I believe that in this Editor's Note, Mr. Browne brought up a very interesting subject—the predominance of papers from the academia rather than the industry at this year's MTT-S. I believe that there is a simple reason for this, the reason being that there isn't much going on these days in the industry as far as invention/research is concerned (compared to academia). One simply cannot invent or do much R&D work on a schedule, especially when the schedule is tight and the company is desperate to meet its financial goals.

The desire to communicate results in a proper environment (such as a conference or symposium) is also more academic than it is entrepreneurial and, in a company's view, the higher purpose of inventing or improving the state-of-the-art is better left to the dreamers or the pioneers. The good news is that, in their strong desire to meet their financial goals (some may call it greed), companies will always use other people's work and results, and they will always pay attention to important academic research, even if the prototypes may not be ready for the production line.

Alex Ciubotaru

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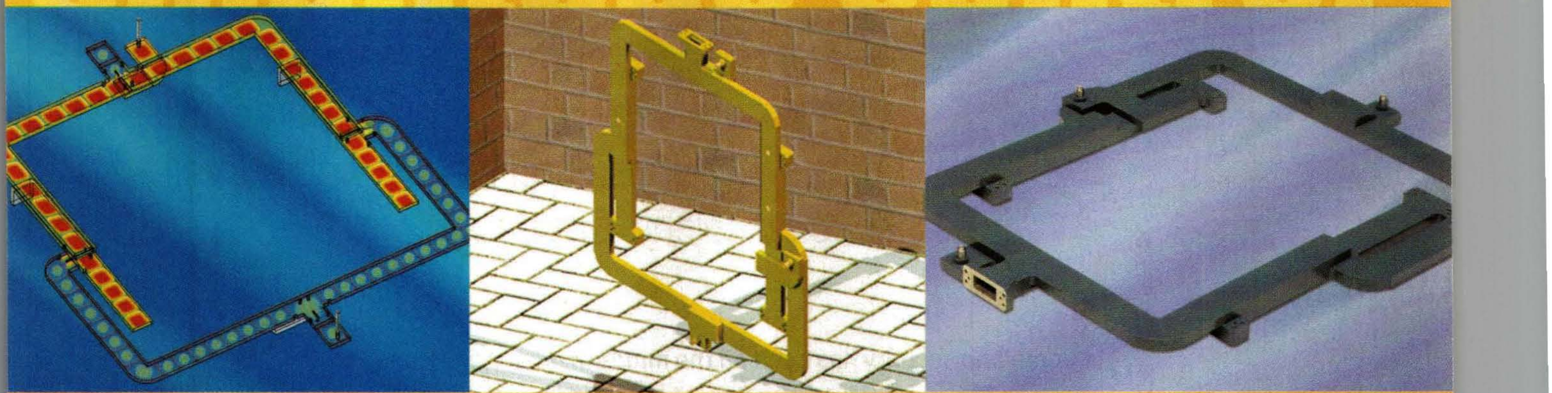

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Meeting The Needs Of Homeland Security

Fear of terrorist attacks on home soil has sparked the growth of a new market for communications, intelligence, and surveillance electronic devices.

JACK BROWNE
Publisher/Editor

Homeland security has been on the minds of most Americans since the terrorist destruction of New York's World Trade Center on September 11, 2001. Although terrorism was a source of fear in Europe and the Far East long before that tragic day, the attack served notice to Americans that they were not immune. The US Department of Homeland Security was established shortly after September 11th, with Tom Ridge, former Governor of Pennsylvania, in the newly appointed Cabinet position as its Secretary. For RF and microwave design engineers, the electronic needs of an evolving homeland security market are many, but not unlike military electronics markets.

Although homeland security is still ill defined as a market, the nature of the electronic equipment required—for communications, intelligence, and surveillance—makes it a natural fit for most defense contractors. It is not surprising to see technological offerings for homeland security from most of the leaders in defense electronics, including BAE Systems, Boeing, Lockheed Martin, Northrop Grumman, Raytheon, and Smiths Industries.

For example, the US Department of Homeland Security (www.dhs.gov) recently awarded a contract extension worth an estimated $198 million to Boeing (www.boeing.com) as part of a contract initiated in October 2003. The contract is part of an effort to improve efficiencies in airport operations and further enhance the aviation security. The award which is managed by the Transportation Security Administration (TSA), is a continuation to the original airport security contract Boeing won in April 2002 to install and maintain explosives detection systems at 429 US airports. In other efforts, Boeing's Connexion is a two-way, broadband satellite communications system capable of linking flight crews with air-traffic control, DHS, and Department of Defense (DoD) personnel.

Integrated Coast Guard Systems (ICGS, Rosslyn, VA), a joint venture between Lockheed Martin (www.lockheedmartin.com) and Northrop Grumman (www.northropgrumman.com) has managed the US Coast Guard's Integrated Deepwater System (IDS) program since it was awarded in June 2002. The multiyear program is designed to modernize and replace the Coast Guard's aging ships and aircraft and improve command and control and logistics systems. This represents the largest acquisition in the history of the Coast Guard, which is responsible for several aspects of homeland security, including maritime security.

Northrop Grumman was recently selected by DHS to participate in the next phase of a program to develop and test anti-missile systems designed to protect commercial aircraft. The company's anti-missile devices are currently deployed on a variety of US and British military aircraft operating worldwide, including C-17 and C-130 military transports. The company hopes to provide effective, economical protection for commercial aircraft by adapting its Directional Infrared Countermeasure (DIRCM) system for this application. Work on the contract will be based at Northrop Grumman's Defensive Systems Div. (Rolling Meadows, IL).

While homeland security issues are generally associated with Federal government programs, considerable investments in homeland security are also made at the state level. M/A-COM, a business unit of Tyco Electronics, has provided multimode public-safety radios for several years based on Internet Protocol (IP) technology. The company's VIDA (voice, interoperability, data, and access) network solution is its newest product line, which uses a combination of voice-over-IP (VoIP) and IP-based digital packet-switched technologies to convert audio and data signals into digital packets. Still, M/A-COM's P25IP digital trunked radio system, NetworkFirst IP packet-switched voice-communications network, and OpenSky radio solution (with 19.2-kb/s data transmission capability) have proven to be cost-effective public-safety communications systems.

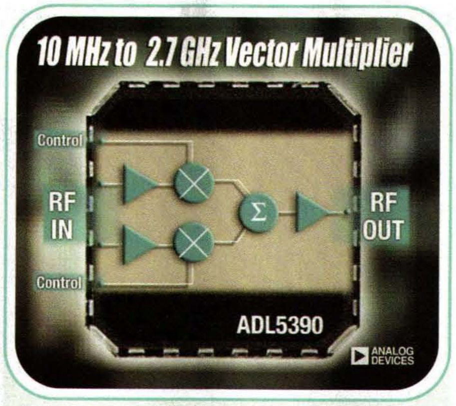
In fact, the OpenSky product won the company a bid to provide statewide public-safety radio system for Pennsylvania in the late 1990s. According to Dr. Dennis Martinez, director of technology for M/A-COM's Wireless Business Unit, proposals were made in the 1998-1999 timeframe, since the state saw a need for a multiple-use communications network. "Users such as the State Police and the Department of Transportation needed to communicate, and have interoperability with other State and Federal agencies," noted Martinez.

More recently, the New York State Office For Technology is in the process of negotiating a 20-year contract worth more than $1 billion to M/A-COM based on the company's bid to replace outdated public-safety radio systems with a digital radio network. As Martinez points out, "We are the prime contractor for the project, in contrast to Pennsylvania, where we are one of three contractors." The radio solution will require a yet-to-be-determined number of towers to be built in the state's Adirondack and Catskill parks for statewide coverage. MRF

Vector Multiplier Controls Gain And Phase

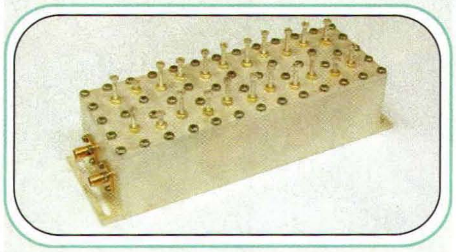
MODEL ADL5390 is the first of its kind, a two-channel vector multiplier that can replace as many as six discrete components previously used for gain and phase control in wireless infrastructure equipment. The new device operates over an input frequency range of 10 MHz to 2.7 GHz and provides a continuous amplitude control range of 35 dB and a phase control range of 360 deg. This new class of component operates on a single +5-VDC supply and achieves an output 1-dB compression point of +13 dBm and an output third-order intercept point of +25 dBm. An output switch disable function allows operators to null RF signals from the output port. The device is supplied in a 24-pin lead-frame chip-scale package (LFCSP) measuring 4 × 4 mm and rated for operating temperatures from –40 to +85°C. P&A: $7.50 (1000 qty.).

Analog Devices, Inc., 804 Woburn St., Wilmington, MA 01887; (781) 937-1989, FAX: (781) 937-1026, Internet: www.analog.com.



ANALOG DEVICES' MODEL ADL5390 VECTOR MULTIPLIER

DOW-KEY'S MODEL M1C06-CDK2 SPDT BIPOLAR LATCHING SWITCH



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MICROELECTROMECHANICAL SYSTEMS (MEMS) are tiny mechanical devices fabricated with semiconductor technology, such as the model M1C06-CDK2 single-pole, double-throw (SPDT) bipolar latching switch. Rated for 100 million switching cycles, the tiny switch uses magnetic actuation to provide high electrical performance from DC to 6 GHz. The switch achieves 45 dB isolation to 3 GHz and 40 dB isolation to 6 GHz, with 0.2 dB insertion loss to 3 GHz and 0.5 dB insertion loss to 6 GHz. The switching time is less than 200 µs and typically less than 50 µs. The SPDT switch exhibits return loss of at least 20 dB to 3 GHz and at least 14.5 dB to 6 GHz. The MEMS switch die measures 1.96 × 1.64 mm and is 5.88 × 5.88 × 2.78 mm in its quasi-hermetic package.

Dow-Key Microwave Corp., 4822 McGrath St., Ventura, CA 93003-7718; (805) 650-0260, FAX: (805) 650-1734, Internet: www.dowkey.com.

Diplexers Separate PCS Uplinks/Downlinks

A LINE OF diplexers is available for maintaining high isolation between Personal Communications Services (PCS) uplink frequencies (1850 to 1910 MHz) and downlink frequencies (1930 to 1990 MHz). Designed to handle input-power levels to 50 W, the diplexer features only 2.0-dB insertion loss in the passband. It is supplied in a compact housing with SMA female connectors and is rated for operating temperatures from –30 to +50°C. Additional models include the XMA-D50-P-20-S with a passband of 20 MHz (2-dB passband insertion loss) and model XMA-D50-P-30-S with a passband of 30 MHz (1.8-dB passband insertion loss).

XMA Corp., 150 Dow St., Manchester, NH 03101; (603) 222-2256, FAX (603) 222-2259, Internet: www.xmacorp.com.

Fast Synthesizers Improve In Reliability

MAJOR DESIGN ENHANCEMENTS have been made to the 2200 series of fast-switching frequency synthesizers. The synthesizers provide +10 dBm output power with ±2-dB output-power flatness over the selected frequency range. The modular architecture of the 2200 series (more than 30 different models in the series with different frequency ranges and frequency resolution) is housed in a 3U-high rack-mount enclosure with front-panel keyboard for manual control and GPIB and binary-coded-decimal (BCD) interfaces for automatic remote control. Ideal for radar system and cross-section measurements, the sources are based on direct-analog frequency-synthesis techniques (mixing, adding, and dividing discrete frequencies) for low phase noise, with –101 dBc/Hz offset 1 kHz from a 2.4-GHz carrier and –128 dBc/Hz offset 10 MHz from the same carrier. P&A: $70,000 and up.

Aeroflex, Inc., 35 South Service Rd., P.O. Box 6022, Plainview, NY 11803-0622; (516) 694-6700, (800) 843-1553, FAX: (516) 694-4823, e-mail: info-test@aeroflex.com, Internet: www.aeroflex.com.


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Record Revenue For Silicon Labs

SILICON LABORATORIES, INC., a firm involved in high-performance, analog-intensive, mixed-signal ICs, has reported record first-quarter revenue for the period ended April 3, 2004. the company announced revenues of $113.6 million in the first quarter, its twelfth consecutive quarter of revenue growth.

Revenue for the first quarter of 2004 increased to $113.6 million from $109.6 million in the fourth quarter of 2003. This represents a 78-percent increase over revenue of $63.8 million during the same period in 2003.

Under generally accepted accounting principles (GAAP), operating income for the first quarter was $27.2 million or 23.9 percent of revenues. First-quarter net income was $19.9 million, resulting in diluted net income per share of $0.36, compared to fourth-quarter 2003 net income of $20.9 million. Excluding non-cash charges for amortization of deferred stock compensation, adjusted net income for the first quarter was $21.2 million, representing 18.6 percent of revenue. Adjusted diluted net income per share was $0.38.

The company ended the quarter with a highly liquid balance sheet with cash and short-term investments totaling $204 million, an increase from $190 million at the end of the fourth quarter of 2003.

"Many of the markets we serve are showing strong demand, including VoIP customer premise equipment, notebook computers, and GSM/GPRS handsets," says Dan Artusi, president and CEO of Silicon Laboratories. "We believe we are leading the competition in both integration and bill-of-materials savings due to a strong portfolio of first-of-a-kind innovations in CMOS. We believe this lead is allowing us to gain significant market share across our key markets."

Silicon Laboratories' revenue growth resulted from solid performance in both the broad-based mixed-signal and mobile handset businesses.

"We expect to experience continued strength across our product lines and further diversification of our customer base," adds Artusi. "We also intend to continue to invest in R&D to further extend our lead and expand our new product pipeline." MRF

CONTRACTS

REMEC Defense & Space Group, MMIC Division—Was awarded a $675,000 contract from the Air Force Research Library (AFRL) Aerospace Components and Subsystems Concepts Division, Sensors Directorate to develop engineering prototype X-band and Ku-band T/R MMICs/modules and power amplifiers (PAs) as low-cost building blocks for future phased-array radar systems that require large quantities of modules.

Vishay Intertechnology, Inc.—Announced that they have contracted Modelithics, Inc. of Tampa, FL for the development of characterization models representative of the high-frequency performance of Vishay's series of silicon-based RF capacitors.

TECOM Industries, Inc.—Announced that AAI Corp. has awarded TECOM a major contract for approximately $2.4 million in support of the Shadow Tactical Unmanned Aerial Vehicle (TUAV) full-rate production program.

TECOM is a major supplier to AAI Corp. for the Shadow TUAV production program, providing small aperture antenna assemblies for Ground Data Terminals (GDTs) as well as dual element C-band and S-band antennas for the aerial vehicle. The GDT antenna assembly consists of a positioner, controller, and parabolic feed assemblies that provide the command, control, and sensor connectivity of the TUAV to the ground control station.

FRESH STARTS

Vodafone—Announced that it has selected Nokia as its 3G network-infrastructure partner.

The main role of Nokia will be to partner Vodafone Australia and New Zealand in the establishment of 3G network infrastructure. In addition to the provision of infrastructure, Nokia is expected to have a key role in network planning, optimization, and operations.

VIDA Products, Inc.—VIDA is a new startup company based in Santa Rosa, CA. They concentrate on high Q filters, low-phase-noise oscillators, and high-performance frequency synthesizers all using YIG technology.

Hittite Microwave Corp.—Has appointed a new sales representative firm to serve customers in France. SALIES SA, headquartered in Grigny, France, was established in 1965 to act as a specialist European representative for electronic-component manufacturers from the US, EU, and Far East. SALIES SA areas of expertise include: RF and microwave components as well as test and measurements equipment including the space, military, and commercial markets.

SALIES can be contacted at +33 1 69 02 25 70, via fax at +33 1 69 02 25 99, or e-mail at ventes@salies.fr.

Laird Technologies—Announced its acquisition of Thermagon, Inc., which is located in Cleveland, OH.

Thermagon's products include thermally conductive silicon pads, phase-change products, adhesives, and putties. In addition to these interface materials, they make thermally conductive epoxy pre-preg (T-preg), core laminates (DSL), and metal-based circuit-board laminates for printed-circuit-board fabricators to manufacture single-layer, double-layer, and multi-layer circuit boards for use in heat dissipation.

QUALCOMM, Inc.—Announced their participation in a $14 million round of venture capital financing in Techfaith Wireless Communication Technology Ltd. ("Techfaith Wireless"), an independent handset design house and China's leading mobile terminal design group. Techfaith Wireless will design handsets based on CDMA2000® technology, initially targeting the Chinese domestic market.

Electri-Flex Co.—Has redesigned its website, which is accessible at www.electriflex.com. Electri-Flex customers and prospects and prospects will have access to the new Conduit Application Guide where they can select the conduit that best suits their needs.

New features on the website include an OEM page with information pertinent to manufacturers who want to incorporate Electri-Flex conduit into their products; literature downloads, which provide information on Electri-Flex Liquatite products and applications; and news and events, which keep readers informed about Electri-Flex's conduit lines and company news.

Trompeter—Announced the signing of a new manufacturer representative: Electronic Marketing Associates, Inc. (EMA). A veteran of more than 30 years in the industry, EMA will be representing Trompeter throughout the Southeastern US from offices in Raleigh, NC, Huntsville, AL, and Atlanta, GA. EMA will cover the states of Alabama, Mississippi, Georgia, and Tennessee.

Spectrum Control, Inc.—Has acquired all of the outstanding common stock of Salisbury Engineering, Inc. (SEI) of Delmar, DE. Salisbury Engineering designs and manufactures a full line of RF and microwave components and systems used primarily in military and aerospace applications.

At this time, Salisbury Engineering employs approximately 40 people at the Delmar, DE facility.

Cree, Inc.—Announced that it has completed the acquisition of the gallium-nitride (GaN) substrate and epitaxy business of Advanced Technology Materials, Inc., a wholly owned subsidiary of ATMI, Inc. for $10.25 million. On March 25, 2004, Cree announced a definitive agreement to purchase the business including related intellectual property, fixed assets, and inventory.

Plexus Corp.—Announced that its board of directors has approved an expenditure of approximately $12 million to expand the company's operations in Penang, Malaysia. The authorization includes the purchase of an existing 164,000-sq.-ft. facility and the initial outfitting of manufacturing equipment. Additionally, Plexus has extended for five years its lease on the 'Building 4' facility located in Neenah, WI. MRF

KENT

AMI Semiconductor Names John Kent To VP Spot

AMI Semiconductor has appointed JOHN KENT as vice president of technology research and development (R&D). In his previous position, Kent was the manager of ASIC and Array development for IBM's Microelectronics Division.

SV Microwave, Inc.—DOUG SNADER to assistant vice president for product and market development; formerly held various management positions at TRU Corp., M/A-COM, and Micro-Coax.

Advantest America, Inc.—DENNIS MALLOY to vice president of sales; formerly vice president of customer operations for Kulicke & Soffa Industries.

APLAC Solutions Corp.—HARRY LILJA to managing director; formerly CEO of Beijer Electronics.

Novellus Systems, Inc.—NEIL R. BONKE to the board of directors; formerly chairman and CEO of Electroglas, Inc. Also, DR. YOUSSEF A. EL-MANSY to the board of directors; formerly vice president and director of logic technology development at Intel Corp.

Somera Communications, Inc.—DAVID W. HEARD to president and CEO; formerly president and general manager of Tekelec, Inc.'s Network Switching Division.

MKS Instruments, Inc.—LEO BERLINGHIERI to president and COO; formerly vice president and COO.

Actel Corp.—DAN MCCRANIE to the board of directors; currently serves as chairman of the board for several semiconductor manufacturers, including ON Semiconductor, Xicor, and Virage Logic.

NEC-Mitsubishi Electronics Display of America, Inc.—DOUG ALBREGTS to vice president of marketing; formerly vice president of North American sales.

Innovex, Inc.—PHILIP D. ANKENY to the board of directors; remains as vice president of business development and CFO at SurModics, Inc.

Alvarion, Inc.—DR. DAVID KETTLER to the board of directors; formerly vice president in charge of the Science and Technology organization at BellSouth.

EMS Wireless—SOREN PIHLMAN to vice president and deputy general manager; formerly president and COO of Digital Transmission Systems, Inc. Also, DAVE MCKAY to CTO; formerly vice president of engineering.

Endicott Interconnect Technologies (EI)—JAMES SULLIVAN to vice president of human resources; formerly employed with Hadco and Sanmina-SCI. Also, LEEANN J. LEVESQUE to general manager for homeland security; formerly director of sales for Sanmina-SCI.

Microtech, Inc.—DON EFFGEN to the position of sales manager; formerly technical sales manager with Eschenbach Optik of America.

Hover-Davis, Inc.—DARRELL R. TULLAR to marketing manager; formerly employed at Motorola, Universal Instruments, and PMJ Automec.

TULLAR

HENCKEL

LEDtronics— JOHN HENCKEL to sales representative for Northern Europe (Denmark, Sweden, Norway, Finland, The Netherlands, Germany, and Austria); formerly employed in the electronics and manufacturing businesses as well as sales to the solid-state lighting arena. MRF

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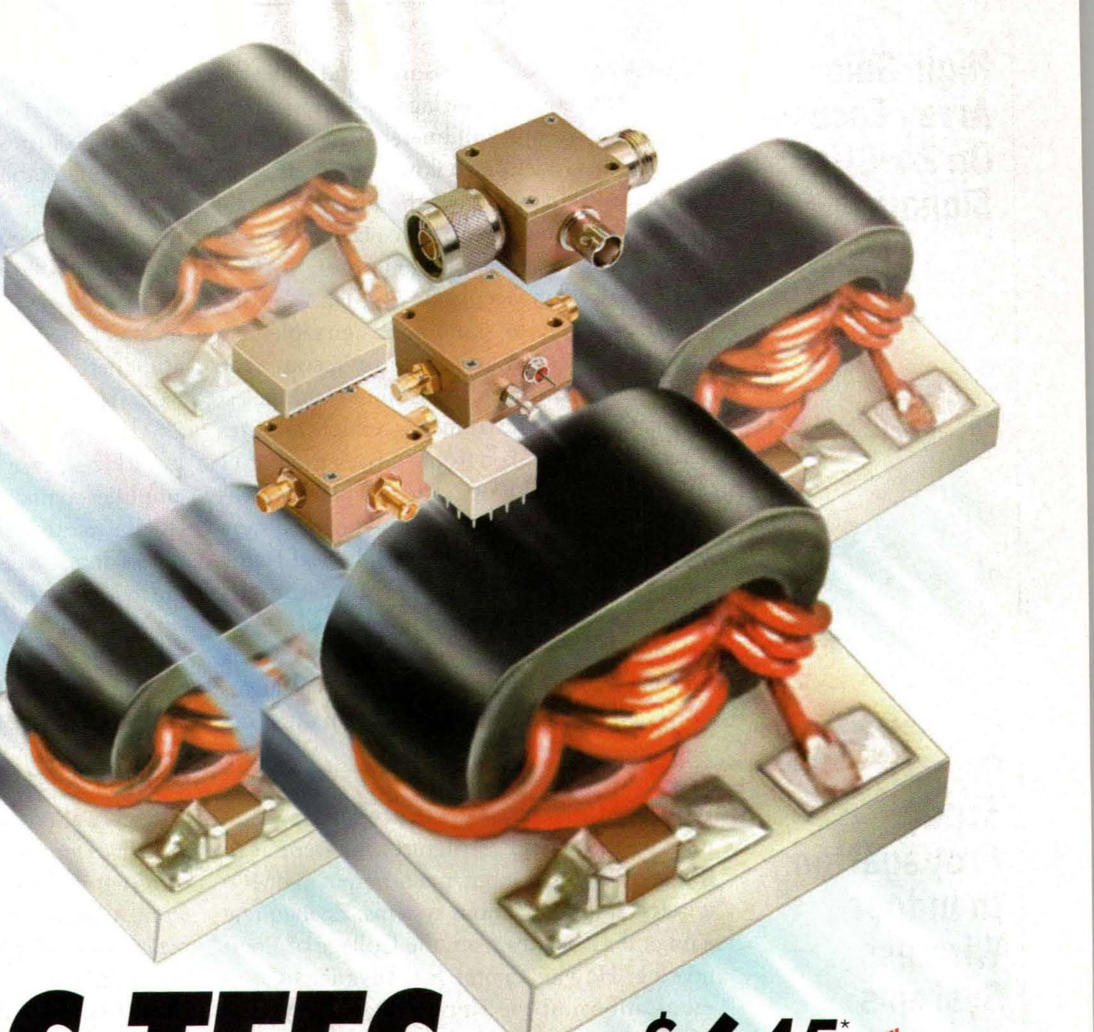

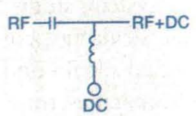

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PBTC-1G	10-1000	0.3	33	1.10	25.95
PBTC-3G	10-3000	0.3	30	1.13	35.95
PBTC-1GW	0.1-1000	0.3	33	1.10	35.95
PBTC-3GW	0.1-3000	0.3	30	1.13	46.95
ZFBT-4R2G	10-4200	0.6	40	1.13	59.95
ZFBT-6G	10-6000	0.6	40	1.13	79.95
ZFBT-4R2GW	0.1-4200	0.6	40	1.13	79.95
ZFBT-6GW	0.1-6000	0.6	40	1.13	89.95
ZFBT-4R2G-FT	10-4200	0.6	N/A	1.13	59.95
ZFBT-6G-FT	10-6000	0.6	N/A	1.13	79.95
ZFBT-4R2GW-FT	0.1-4200	0.6	N/A	1.13	79.95
ZFBT-6GW-FT	0.1-6000	0.6	N/A	1.13	89.95
ZNBT-60-1W	2.5-6000	0.6	45	1.10	82.95

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High-Gain Array Focuses On 24-GHz Signals

PLANAR ANTENNA DESIGNS are useful for a variety of applications requiring small physical size at microwave and millimeter-wave frequencies. Phillip Grajek and fellow researchers at the Radiation Laboratory of the Electrical Engineering and Computer Science Department at the University of Michigan (Ann Arbor, MI) have developed a compact planar 24-GHz Yagi-Uda antenna array using standard design tables and scaling techniques to take into account the added capacitance due to the supporting dielectric substrate. The antenna, which features directivity of 9.3 dB, a front-to-back ratio of 11 dB, and bandwidth of 2.5 to 3.0 percent, was implemented as part of an 11-beam system using a planar array and a 2-in. spherical Teflon lens. Measured beam patterns show beam gain of 22 dB with cross-polarization level of –24 dB and a crossover level of –6 dB. The Yagi-Uda antenna employs two forward director elements and one reflector to maximize beam directivity. It was fabricated on 10-mm-thick Duroid substrate material with dielectric constant of 2.2. The initial dimensions of the table were obtained from design tables for maximum directivity in an air dielectric using two directors, one reflector, and cylindrical-wire elements. The Yagi-Uda antenna was fabricated using standard lithography and wet etching techniques. The antenna was tested in receive mode using a lock-in amplifier and a planar zero-bias Schottky diode from Aeroflex/Metelics as the RF detector. The 24-GHz source was amplitude modulated with a 1-kHz square wave to enable the lock-in amplifier to make measurements on beam levels. The researchers note that the scaling approach to this antenna design can readily be applied to much higher frequencies for use in millimeter-wave applications. See "A 24-GHz High-Gain Yagi-Uda Antenna Array," *IEEE Transactions on Antennas and Propagation,* May 2004, Vol. 52, No. 5, p. 1257.

Study Propagation In Indoor Wireless Systems

UNDERSTANDING PROPAGATION effects and fading due to complex structures is essential to the design of multi-input, multi-output (MIMO) wireless communications systems. Zhengquing Yun and associates from the College of Engineering, Hawaii Center for Advanced Communications, at the University of Hawaii at Manoa (Honolulu, HI) have investigated the influence of complex wall structures on indoor wireless communications using a two-dimensional finite-difference time-domain (FDTD) method to model the electric field distributions. Two cases were studied. The first considered all walls to be homogenous solid slabs. The second assumed that the walls were more complex structures of varying densities. In their studies, the researchers discovered that the Rician K factors for the two cases could differ by as much as 5 dB. This difference also resulted in quite different MIMO capacities for wireless systems studied under the different conditions, deviating considerably from formerly assumed ideal conditions. To measure the MIMO capacities, transmit antennas were fixed while receive antennas were moved along three directions, resulting in 350 distributed test locations. In the case of both models and measurements, a frequency of 900 MHz was used. with a relative permittivity of 3 for the homogenous walls. The researchers found that capacity diminished with wall complexity, although more studies were needed. See "Complex-Wall Effect on Propagation Characteristics and MIMO Capacities for an Indoor Wireless Communication Environment," *IEEE Transactions on Antennas and Propagation,* April 2004, Vol. 52, No. 4, p. 914.

Low-Power CMOS TIA Handles Signals To 20 GHz

TRANSIMPEDANCE AMPLIFIERS (TIAS) are essential building blocks in high-speed optical communications systems. The TIA converts current from a photodiode into an amplified voltage. Most TIAs with bandwidths exceeding 10 GHz have been fabricated in GaAs, InP, or SiGe. Christian Kromer and a team of researchers from the Swiss Federal Institute of Technology (Zurich, Switzerland), however, have developed a low-power 20-GHz TIA in 80-nm silicon CMOS process technology capable of 52.8dB transimpedance gain. The measured TIA bandwidth with a photodiode is 13.4 GHz and without the photodiode 22.6 GHz. The power consumption at 1 V is a mere 2.2 mW. The noise level of the device with reference to the input is a very low 28 pA/√Hz, while the sensitivity at 20 GHz is better than –8 dBm. The TIA features a common-gate topology with a gain-enhancing feedforward path and input-impedance-reducing feedback. See "A Low-Power 20-GHz 52-dBΩ Transimpedance Amplifier in 80-nm CMOS," *IEEE Journal of Solid-State Circuits*, June 2004, Vol. 39, No. 6, p. 885.

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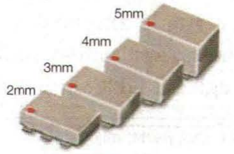

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ADE-3L	+3	0.2-400	5.3	47	10	4	4.25
ADEX-10L	+4	10-1000	7.2	60	16	3	2.95
ADE-1	+7	0.5-500	5.0	55	15	4	1.99▲
ADE-1ASK	+7	2-600	5.3	50	16	3	3.95
ADE-2	+7	5-1000	6.67	47	20	3	1.99▲
ADE-2ASK	+7	1-1000	5.4	45	12	3	4.25
ADE-6	+7	0.05-250	4.6	40	10	5	4.95
ADEX-10	+7	10-1000	6.8	60	16	3	2.95
ADE-12	+7	50-1000	7.0	35	17	2	2.95
ADE-4	+7	200-1000	6.8	53	15	3	4.25
ADE-14	+7	800-1000	7.4	32	17	2	3.25
ADE-901	+7	800-1000	5.9	32	13	3	2.95
ADE-5	+7	5-1500	6.6	40	15	3	3.45
ADE-5X	+7	5-1500	6.2	33	8	3	2.95
ADE-13	+7	50-1600	8.1	40	11	2	3.10
ADE-11X	+7	10-2000	7.1	36	9	3	1.99▲
ADE-20	+7	1500-2000	5.4	31	14	3	4.95
ADE-18	+7	1700-2500	4.9	27	10	3	3.45
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ADE-3G	+7	2300-2700	5.6	36	13	3	3.45
ADE-28	+7	1500-2800	5.1	30	8	3	5.95
ADE-30	+7	200-3000	4.5	35	14	3	6.95
ADE-32	+7	2500-3200	5.4	29	15	3	6.95
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ADE-1MHW	+13	0.5-600	5.2	53	17	4	6.45
ADE-10MH	+13	800-1000	7.0	34	26	4	6.95
ADE-12MH	+13	10-1200	6.3	45	22	3	6.45
ADE-25MH	+13	5-2500	6.9	34	18	3	6.95
ADE-35MH	+13	5-3500	6.9	33	18	3	9.95
ADE-42MH	+13	5-4200	7.5	29	17	3	14.95
ADE-1H	+17	0.5-500	5.3	52	23	4	4.95
ADE-1HW	+17	5-750	6.0	48	26	3	6.45
ADEX-10H	+17	10-1000	7.0	55	22	3	3.45
ADE-10H	+17	400-1000	7.0	39	30	3	7.95
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Antenna Snares GPS/WLAN Signals

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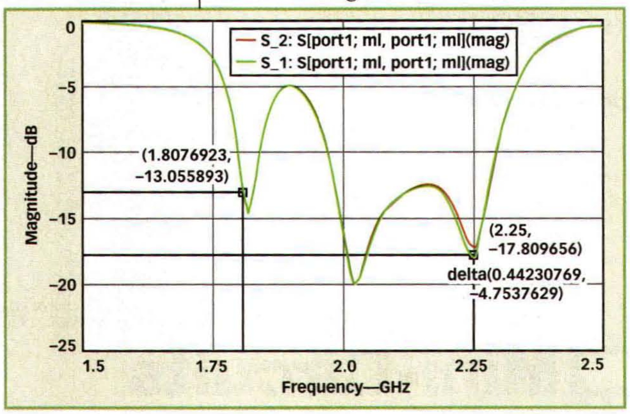
High gain is not usually associated with a compact antenna. For satellite-communications applications, however, an antenna design must be small and light, yet provide pattern shaping, wide bandwidth, and polarization purity. In developing an antenna for multiband Global Positioning System (GPS) and wireless-local-area-network (WLAN) use, it was possible to create a small, lightweight design with polar-ization diversity and high gain.

For GPS applications, for example, an antenna may be required to handle both the low band at 1.226 GHz and the high band at 1.575 GHz. For IEEE 802.11a/b/g WLANs, an antenna must operate in both the 2.4- and 5-GHz bands with bandwidth to support data rates at 11 and 54 Mb/s. Additional applications include the planned Air Force satellite systems at 1.8 and 2.25 GHz. For a single antenna to cover multiple wireless applications, coverage at 1.8 through 2.1 GHz should also be considered for third-generation (3G) cellular systems.



1. This simulation of antenna return loss shows two desired resonances at 1.8 and 2.25 GHz as well as a third resonance at about 2.1 GHz. The simulation indicates a design that can be used as a triple-band antenna or a wideband antenna.

Polarization is an important characteristic in a good antenna design. For space applications, it is customary to use circular polarization (CP), such as the right-handed circular polarization (RHCP) or left-handed circular polarization (LHCP), either for transmit and receive or reuse of the same spectrum band for added capacity. Although most WLAN systems require linear polarization, the use of CP will eventually becomes advantageous for mobile systems.

Certain theoretical limits dictate how small an antenna can be made while still providing required gain and bandwidth. For the space-based (satellite) application, the antenna design was required to fit a certain form factor while operating with CP at 1.8 GHz for the uplink (the satellite's receive frequency) and 2.25 GHz for the downlink (the satellite's transmit frequency). Pattern-shaping capability was also a key requirement, to allow the satellite to maintain communication while in different positions and angles. The antenna must be rugged enough to withstand high shock and vibration,

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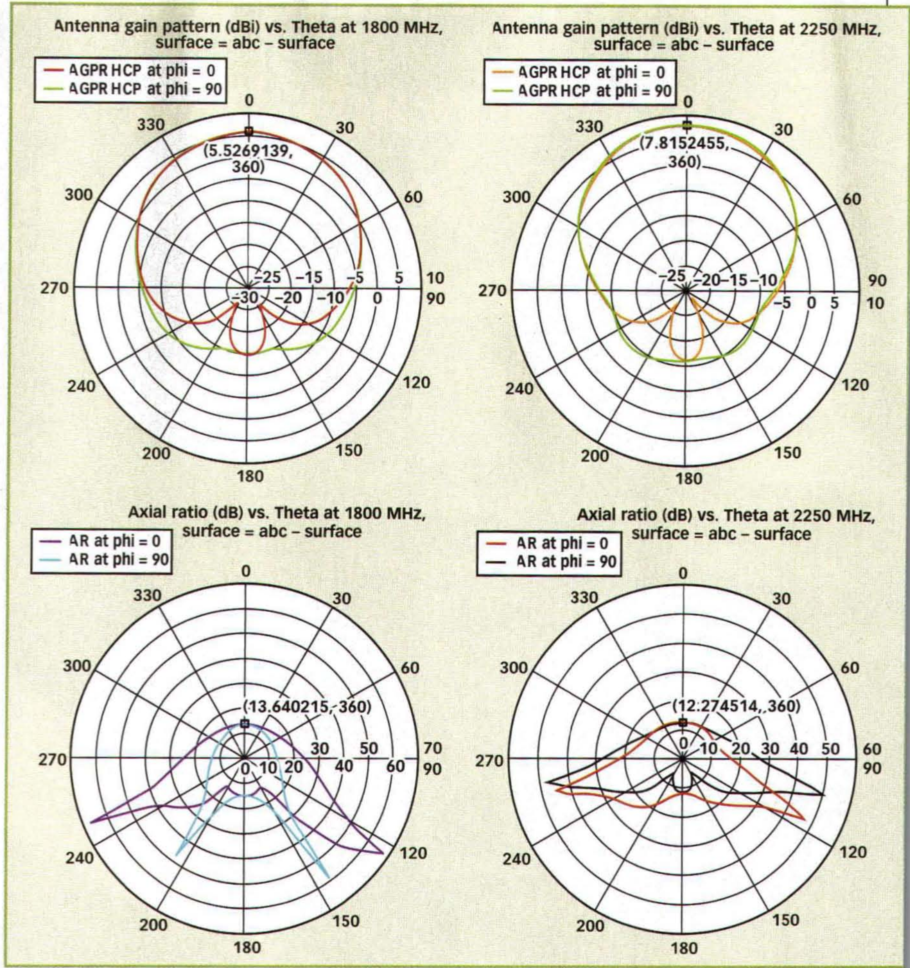

2. These simulations show the radiation pattern (top left) for the antenna of Fig. 1 at 1.8 GHz along with its axial ratio (bottom left) as well as the radiation pattern at 2.25 GHz (top right) along with its axial ratio at that frequency (bottom right). At the zenith at 1.8 GHz, the predicted RHCP gain is 5.5 dBic with 13.6 dB axial ratio while at 2.25 GHz, the RHCP gain is 7.8 dBic with axial ratio of 12.2 dB.

environmental/thermal extremes (temperature swings of typically –40 to +70°C), and power cycling. Several options were considered for the design, including a helical antenna, quadrifilar helical antennas (QFHAs), and various microstrip patch configurations. Initial analysis and electromagnetic (EM) software simulations indicated the difficulty of achieving the required performance levels in the small physical size.

After considering several unconventional approaches, a ring radiator technology was selected as the potential solution. This approach uses the resonant structures to effectively create long paths (and high gain) for the radiation currents while reducing antenna size by 25 to 35 percent compared to other approaches. The technique made it possible to meet the form-factor requirements with higher gain than possible with an even larger microstrip patch antenna or a cavity-backed helical antenna.

Design and analysis of ring antennas requires very intuitive engineering (and educated guessing) compared to the better-understood design and analysis approaches used for microstrip patch antennas. Fortunately, by performing a detailed initial design and analysis procedure, and carefully reviewing the EM simulation results, it was possible to reduce the design risk for the ring antenna in spite of its complexity.

In a simple rectangular patch antenna, the source of radiation can be thought of as two slots at the two ends of the

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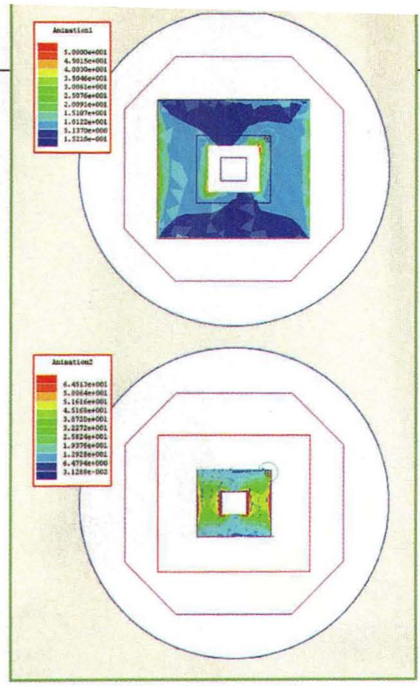
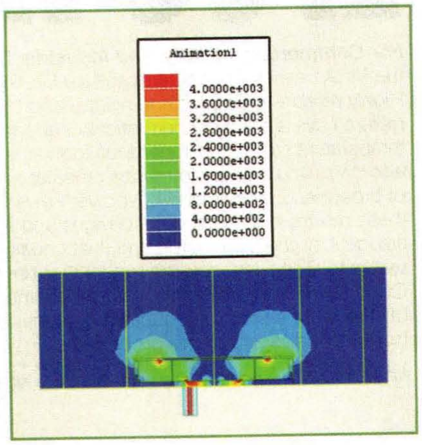
patch, approximately one-half wavelength apart. If each of these slots is about one-half wavelength in length, 2.1 dBi gain should result. Any two such antennas working as a two element array should theoretically provide an additional 3 dB gain. Therefore, it should be possible to achieve about 5.1-dBi gain from a simple patch antenna. With some refinement, it may be possible to get even better gain or pattern shaping depending on the ground plane type or resonance mode.

For a ring antenna, it may be possible to design a structure with multiple resonances that could be spaced and coupled for multiple-frequency or broadband-frequency use.

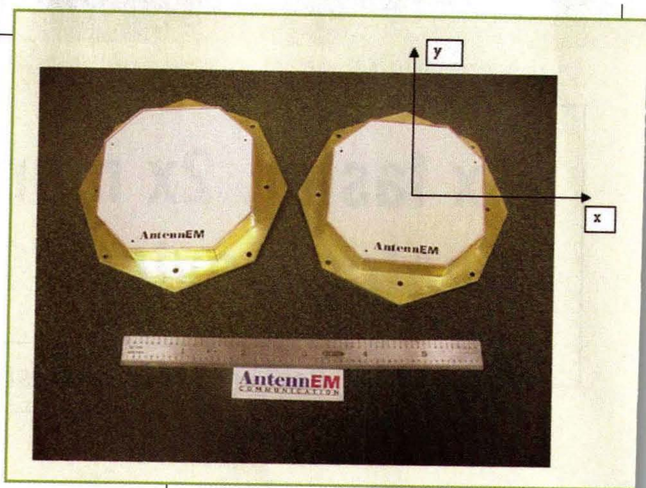


3. This simulation of surface current density shows the upper (top) and lower rings, with the greatest concentration of current on the outside edges.

High gain and pattern shaping can be achieved by properly phasing the various modes to operate in a predetermined manner so as to combine constructively or destructively in the far field and in appropriate directions. In most cases, it was possible to achieve as much as 9 dBic gain (theoretically) from these structures with as much as 17 percent bandwidth. In theory, it was possible to achieve bandwidths of 15, 20, and 30 percent for VSWRs of 1.50:1, 2.0:1, and 3.0:1, respectively. Unfor-



4. This side view simulation of the antenna shows the current hot spots on the edges at 2.25 GHz.



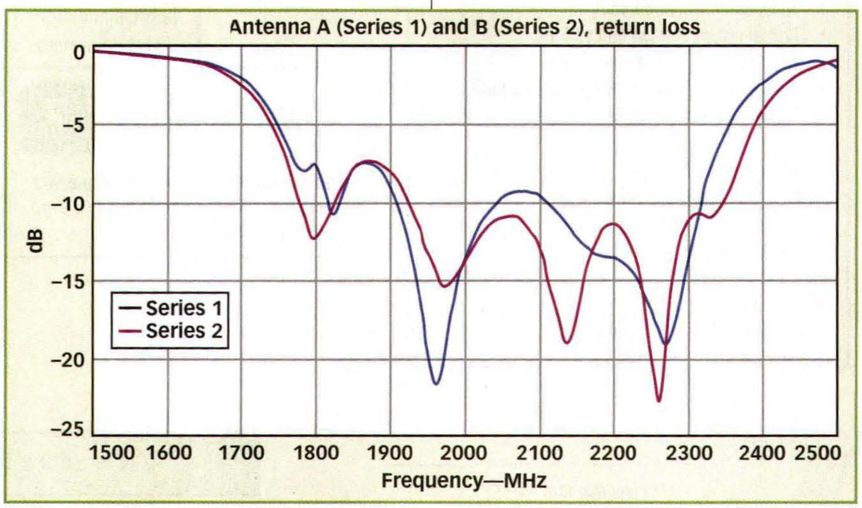
5. These prototype antennas A (left) and B (right) were built according to intuitive design practices and thorough simulation and analysis. Zenith is along the z-axis (coming through the page); the x-axis (to the right) and y-axis are shown for reference.

tunately, it was not possible to develop a systematic design approach that would meet all required physical and electrical parameters at all frequencies. With some effort, it may be possible to develop an approach that would meet the requirements for certain modes of operation, however.

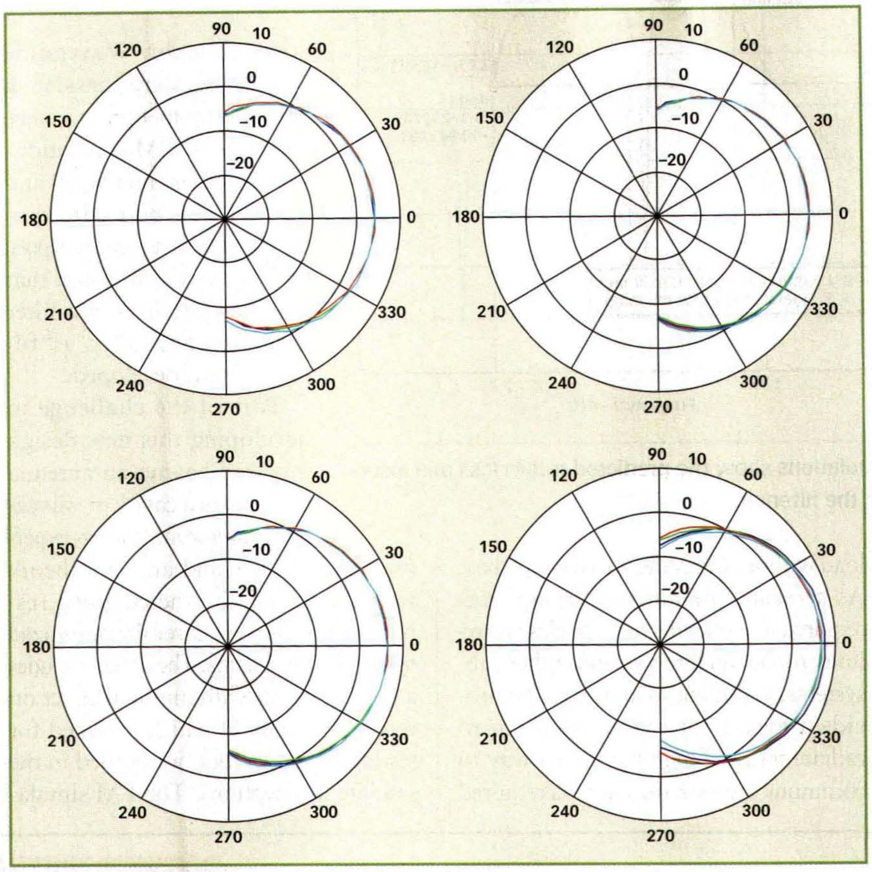
Figure 1 shows a swept-frequency plot as it was predicted by EM simulation for an optimized antenna design. The plot indicates multiple resonances, although not all would be used for the satellite antenna The lowest resonance at 1.8 GHz provides better than 13-dB return loss, while at the high end, at 2.25 GHz, the return loss can be better than 17 dB. For a certain combination of parameters, it is possible to achieve a 10-dB return-loss bandwidth of about 15 percent. This would be an excellent broadband antenna for many applications. The return loss for the resonance at 2.1 GHz is even better, at nearly 20 dB. The antenna design's multiple resonances make it suitable for use as a single wideband antenna or for three discrete-frequency applications.

The predicted radiation pattern for the RHCP antenna at the lower resonance of 1.8 GHz indicates nearly 5.5 dBic gain (**Fig. 2**, top left). The axial ratio is nearly 13 dB at zenith (Fig. 2, bottom left). At the upper resonance of 2.25 GHz, the predicted radiation pattern indicates nearly 8-dBic gain (Fig. 2, top right). At this frequency, the axial ratio appears to be nearly 12 dB (Fig. 2, bottom right).

Figure 3 shows simulations of the surface current densities on the rings. As expected, the highest current densities (in red)—indicating the radiation mechanism for this configuration—can be found along the edges. The top illustration simulates the upper ring while the bottom illustration shows a simulation of the lower ring for the upper resonance at 2.25 GHz. The radiation mechanism changes slightly at the lower resonance, where the gain is lower, although this



6. These measurements show the return-loss performance of the two antenna designs, comparing favorably to the simulated data of Fig. 1.



7. These measurements show the radiation pattern of antenna A at 1.8 GHz, with gain of 4.8 dBic (top left) and at 2.25 GHz, with gain of 7.4 dBic (bottom left), along with the radiation pattern of antenna B, with gain of 4.8 dBic at 1.8 GHz (top right) and 7.8 dBic at 2.25 GHz (lower right). In both cases, azimuth cuts are at 0, 45, 90, and 135 deg. For both antennas, the measured axial ratio was about 12 and 10 dB for each band, respectively.

was optimized according to the requirements of the satellite link budget.

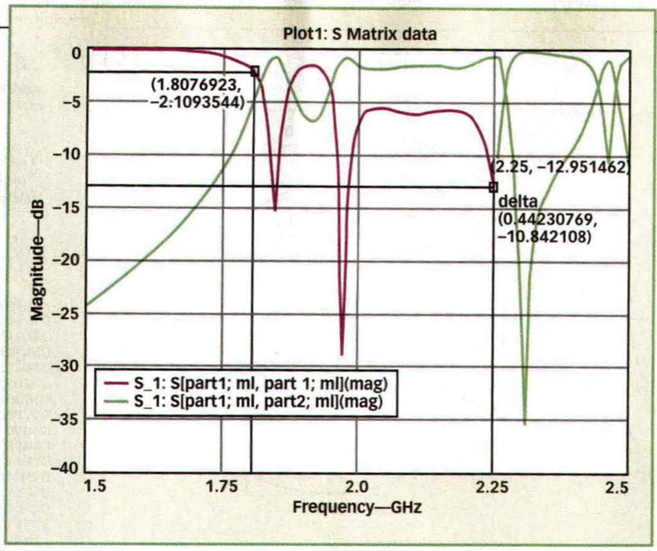
From a side view **(Fig. 4)**, the radiation mechanism can be seen with the coaxial input connector. The larger box surrounding the antenna is a convention of the EM simulation program in which the device to be modeled is enclosed within finite boundaries (the box). The enclosure's boundaries were chosen to have minimal effects on the antenna's performance.

Based on these analyses and simulations, several antennas were manufactured, with two such designs shown in **Fig. 5** (antenna A on the left and antenna B on the right). The antennas met all the electrical requirements and passed all the space qualification requirements the first time around due largely to a well-controlled design process, extensive use of simulation and verification, and good mechanical design and fabrication practices.

Figure 6 shows the measured return loss of antennas A and B, with response shapes remarkably similar to the desired simulated performance of Fig. 1. Differences between the simulations and actual hardware can be traced to some tuning performed in the laboratory, albeit

8. The "filtenna" combines the functionality of a filter and an antenna. The hat section is used for testing and not required for satellite use.

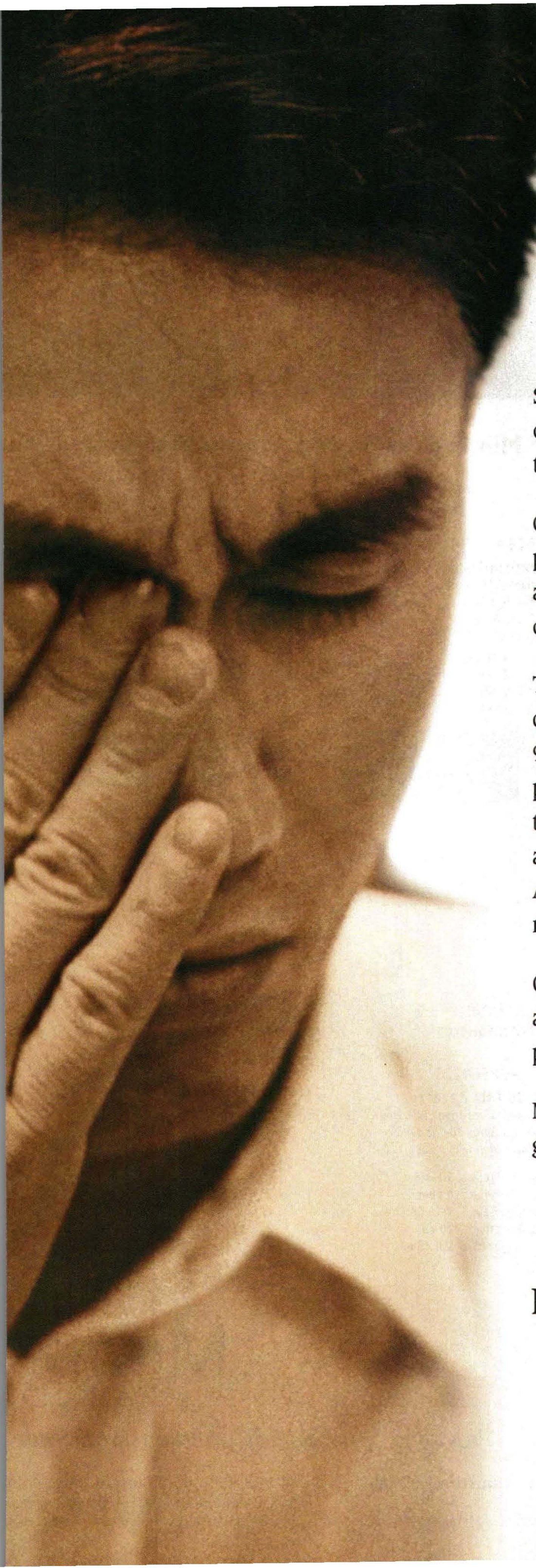
minor. The measured radiation patterns and gains for the two antennas are shown in **Fig. 7**, with measurements at 1.8 and 2.25 GHz for antenna A in Fig. 7 left top and left bottom, respectively, and at 1.8 and 2.25 GHz for antenna B in Fig. 7 right top and right bottom, respectively. Each pattern includes 0, 45, 90, and 135 deg. azimuth pattern cuts. Note the similarity of these measured patterns to the simulations of Fig. 2. The measured backlobe performance was similar to the simulations but not measured for all the antennas.

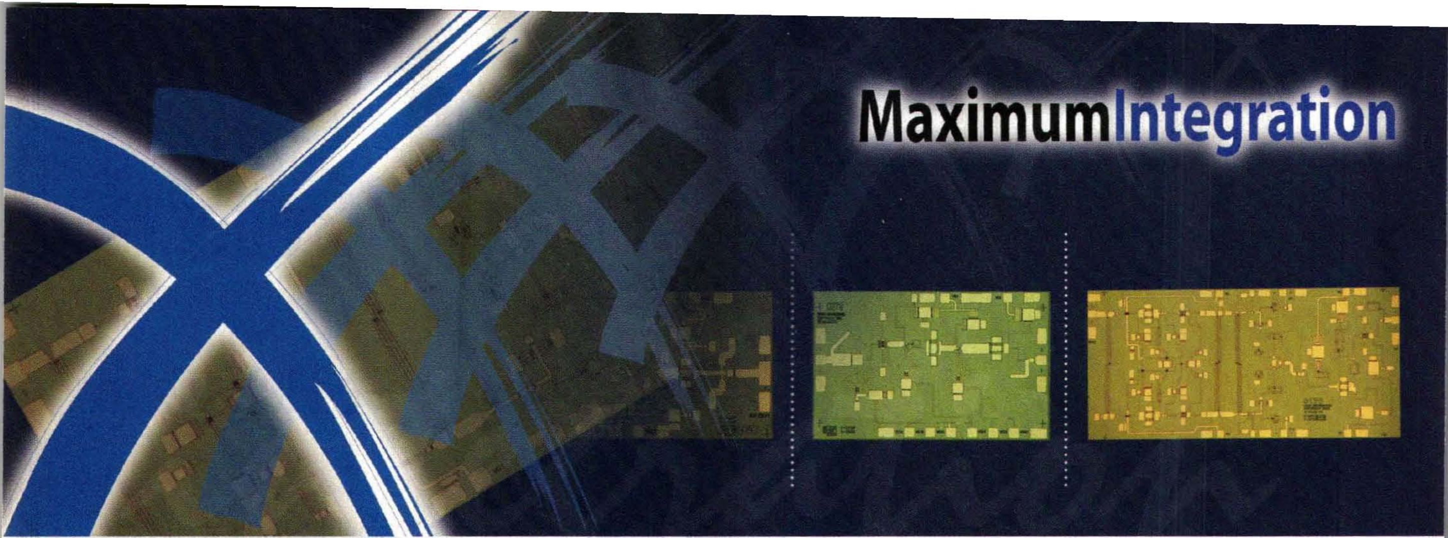


9. These simulations show the predicted return loss and insertion loss for the filtenna.

In addition to the "conventional" antenna requirements, prelaunch ground testing of the satellite payload required a way to test the communications link in the satellite's high bay and to provide communication with the satellite payload without radiating into the high bay. As a result, the antenna design was required to work effectively in close proximity to the satellite's various other subsystems, including solar arrays. To provide a means for the antenna not to radiate into the high bay and a way to communicate with the antenna required some thought. Waveguide approaches were considered, but the form factors were impractical. EM simulations of various antenna boxes and "hats" were performed to identify cutoff behavior and hotspots, and ultimately, a design that combined aspects of a filter and antenna, called a "filtenna," was developed.

Part of the challenge in developing this new design involved having an antenna resonate in a cavity or waveguide. After some fruitless experimentation, filter and antenna theory were combined and the coupling resonator models were carefully optimized to develop the filtenna. The design includes a hat-like cover with minimal effect on return loss **(Fig. 8)** which is added for testing purposes (but not needed in the satellite application). The EM simula-





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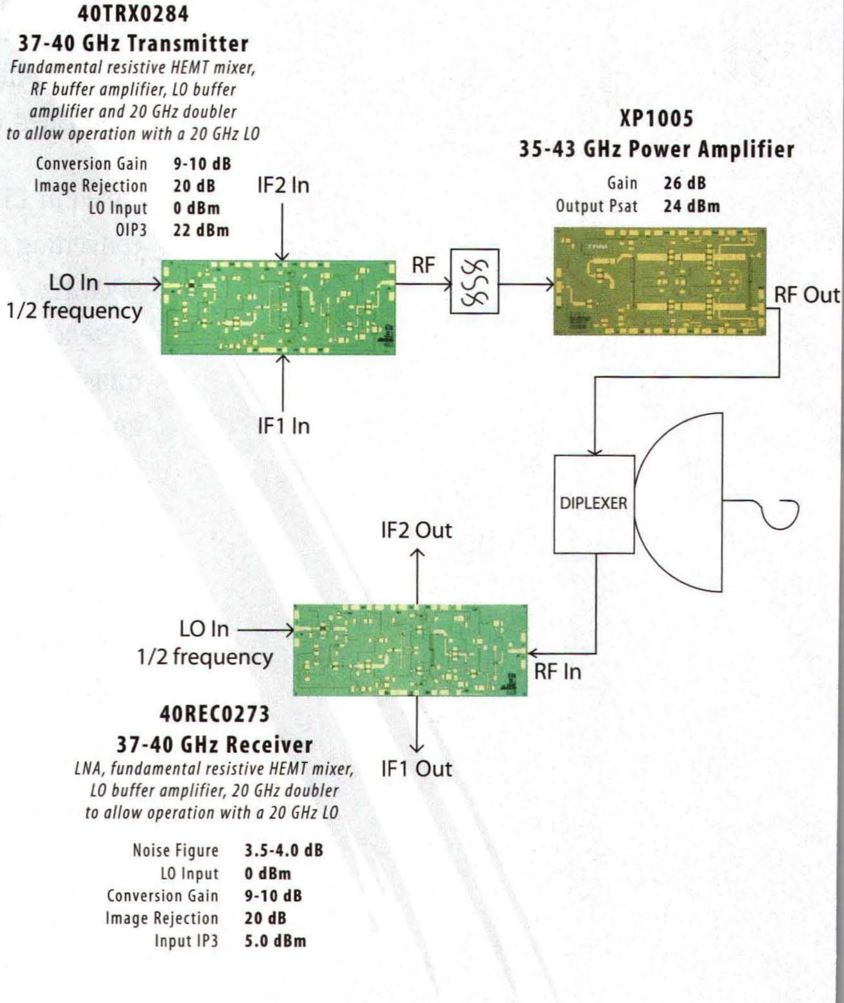

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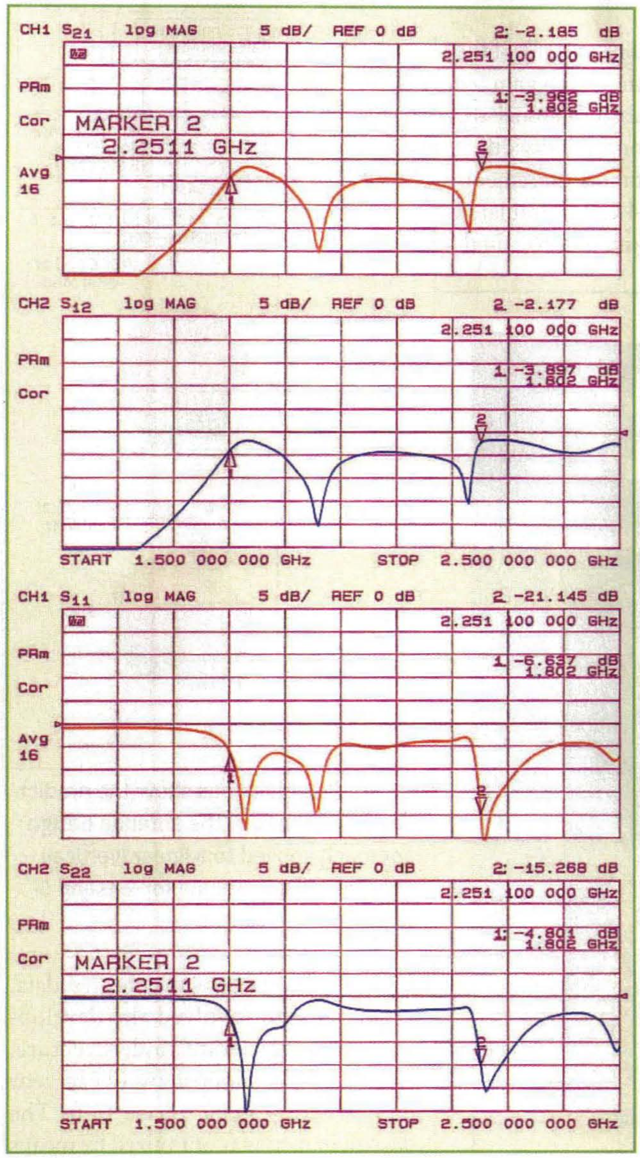
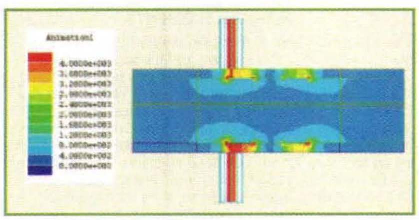


10. These plots show the measured two-port insertion loss (top) and two-port return loss (bottom) for the filtenna.

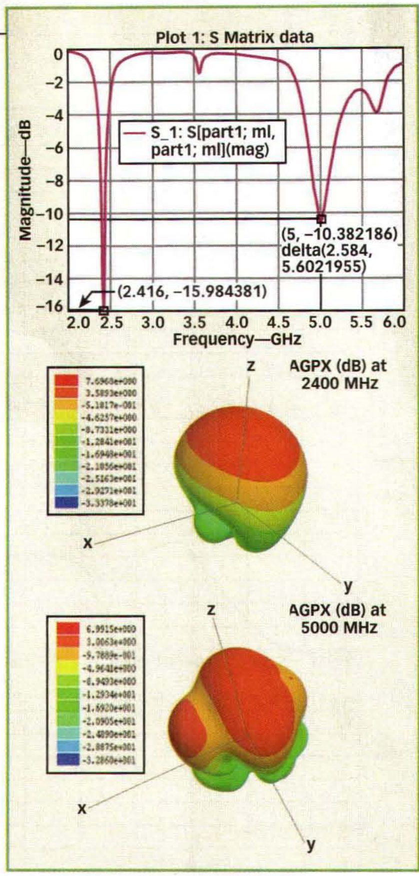
tions showed that the position of the resonances, particularly lower resonances, was very sensitive and shifted with the position of the hat/cover. Simulated results of return loss and insertion loss are shown in **Fig. 9**, while **Fig. 10** shows measured two-port insertion loss (top) and two-port return loss (bottom). The simulated and measured data agree closely, with the exception that the measurements including some tuning in the laboratory to improve the lower bandedge return loss. **Figure 11** shows a side view of the simulated EM fields of the filtenna,



11. This EM field simulation of the filtenna shows the coupling mechanism from the input to the output.

indicating the mechanism of coupling from one port to the other.

This design was applied to two other applications, one for a dual-band Wi-Fi antenna suitable for IEEE 802.11a/b/g WLAN "hot-spot" applications at 2.4 and 5 GHz and the other for dual-band GPS use. **Figure 12** shows simulations of the Wi-Fi antenna, indicating high gain for the linearly polarized design, although the design requires additional bandwidth at the lower end to meet the requirements of IEEE 802.11g at 2.4 GHz. Simulated performance of the antenna for dual-band GPS (not shown) matches test data.



12. These simulations show the predicted performance of the antenna design approach applied to a linear (vertical) polarized Wi-Fi antenna for 2.4- and 5-GHz bands.

The design involved the development of a degenerate mode structure, which would support two very near modes with 90-deg. phase shift. The design, in fact, was optimized by means of this. Still, it would have been helpful to have been able to map the antenna's fields after prototyping to check magnitudes and phases. By optically mapping the field vectors and comparing them with simulated results, it would then be possible to tune the phases of the modes. Such a tool would further reduce the guesswork from the engineering of antennas; it is currently available but as of yet too expensive for practical design. MRF

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Refined I/Q Imbalance Measurements

A fresh approach is needed for the accurate characterization of the analog I/Q modulators and demodulators used in mobile radios with complex modulation.

Phase modulation is the basis for many modern mobile telecommunications formats, including IS-136 and GSM. Using in-phase (I) and quadrature (Q) signal components, modulation is commonly generated by means of an analog I/Q modulator. Since variations in components and manufacturing processes can easily introduce imbalance in I/Q modulator arms, resulting in poor transmission quality, I/Q imbalance measurements are critical to maintaining good system performance. This report will examine such measurements and highlight why traditional measurements can be misleading, offering an improved, alternative approach which can be applied to existing test processes.

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Figure 1 shows a model for data transfer in a digital wireless-communication system. The particular modulation format is not critical here; the important assumption is that the data modulator creates complex signals with I and Q components. The I and Q components can be treated as a complex number, with I being the real part and Q being the imaginary part of the complex number. As a convention, complex quantities will be underlined throughout for clarity.

Since a complex signal cannot be transmitted, it must be converted to a real value and modulated onto a carrier signal. The I/Q modulator provides the means for this. An I/Q modulator consists of a local oscillator (LO) feeding two (I and Q) mixers. The LO signal that feeds the Q mixer is shifted by $+\pi/2$ rad. in phase relative to the LO signal that feeds the I mixer. The mixers' outputs are combined to form the complex signal.

Due to differential additional α_I and α_Q phase shifts along the path leading to the mixer, the phase shift might not be exactly $+\pi/2$ rad., however. Due to different gains, G_I and G_Q, the two mixers may also receive LO signals at different levels, further complicating matters. These errors can be expressed by the complex gains of Eqs. 1 and 2.

Due to leakage, some portion of the LO signal may also appear at the output of the mixers. Such leakage can stem from various sources, which are not discussed here, but it can be represented by additive offsets, O_I and O_Q, to the in-phase and quadrature-phase input signals, respectively. These errors are traditionally characterized by the measures of Eqs. 3-5. Ideally, Eqs. 3 and 4 should be 0 dB.

In Eq. 5, $\underline{O}$ is called the "complex origin offset" and is shorthand for the expression in Eq. 6.

In Eq. 5, $P_{\underline{s}}$ is the average power of the baseband input signal $\underline{s}(t)$. Ideally, this measure should be as low as possible.


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$$\underline{G}_I = G_I\, e^{j\alpha_I} \quad (1) \qquad \underline{G}_Q = G_Q\, e^{j\alpha_Q} \quad (2)$$

$$Gain\,\mathrm{Im}\,balance = 20\ \log_{10}\left|\frac{\underline{G}_Q}{\underline{G}_I}\right| \quad (3) \qquad QuadratureSkew = \ <\left(\frac{\underline{G}_Q}{\underline{G}_I}\right) \quad (4)$$

$$LOLeakage = 10\log_{10}\left(\frac{|\underline{O}|^2}{\left|\frac{\underline{G}_I + \underline{G}_Q}{2}\right|^2 P_{\underline{s}}}\right) \quad (5)$$

$$\underline{O} = O_I\underline{G}_I + jO_Q\underline{G}_Q \quad (6)$$

$$\underline{r}(t) = \underline{G}_I\,\mathrm{Re}\{\underline{s}(t)\} + j\underline{G}_Q\,\mathrm{Im}\{\underline{s}(t)\} + \underline{O} \quad (7)$$

$$\underline{s}(t) = e^{j\delta}\,\underline{s}'(t) \quad (8) \qquad \underline{r}(t) = \underline{G}'_I\,\mathrm{Re}\{\underline{s}'(t)\} + j\underline{G}'_Q\,\mathrm{Im}\{\underline{s}'(t)\} + \underline{O}' \quad (9)$$

$$\underline{r}(t) = \begin{pmatrix} (\underline{G}_I\ \cos\delta + j\underline{G}_Q\sin\delta)\mathrm{Re}\{\underline{s}'(t_k)\} + \\ j(\underline{G}_Q\ \cos\delta + j\underline{G}_I\ \sin\delta)\mathrm{Im}\{\underline{s}'(t_k)\} + \underline{O}' \end{pmatrix} \quad (10)$$

$$\underline{G}'_Q = \underline{G}_I\ \cos\delta + j\underline{G}_I\ \sin\delta \quad (11)$$

$$\underline{G}'_Q = \underline{G}_Q\ \cos\delta + j\underline{G}_I\ \sin\delta \quad (12) \qquad \underline{O}' = \underline{O} \quad (13)$$

$$Gain\,\mathrm{Im}\,balance' = 20\log_{10}\left|\frac{10^{\frac{Gain\,\mathrm{Im}\,balance}{20}}\,e^{jQuadratureSkew} + j\ \tan\delta}{1 + j10^{\frac{Gain\,\mathrm{Im}\,balance}{20}}\,e^{jQuadratureSkew}\ \tan\delta}\right| \quad (14)$$

In practice, this is –30 dB or lower.

The RF channel incorporates all the RF components between the I/Q modulator and I/Q demodulator. As a simple but usable model (especially in manufacturing where the mobile units are connected to test equipment by means of a cable), this signal path is considered to be free of distortion. The signal path is assumed without loss of generality to have 0 dB gain and 0 group delay. As part of the RF signal chain, the I/Q demodulator, which translates a received signal back to complex baseband signals, is also considered to be ideal for modelling purposes. Additional assumptions are that the LO for the I/Q modulator has unit amplitude while the LO for the I/Q demodulator has an amplitude of two units, and that both LOs have a starting phase of 0 when time, t, is 0.

In this model, the received complex baseband signal, $\underline{r}(t)$, can be described by Eq. 7. The $\mathrm{Re}\{\underline{s}(t)\}$ and $\mathrm{Im}\{\underline{s}(t)\}$ operators return the real (I) and imaginary (Q) components of the baseband input signal, $\underline{s}(t)$, respectively.

Generally, the aim of an I/Q modulator characterization is to measure gain imbalance, quadrature skew, and LO leakage since these parameters are used in I/Q modulator data sheets, and because manufacturers often provide some adjustable devices (e.g., potentiometers) that are dedicated to reduce these errors. Equation 7 provides a way to determine the imbalances via estimation of the $\underline{G}_I$, $\underline{G}_Q$, and $\underline{O}$ intermediate parameters. Once $\underline{G}_I$, $\underline{G}_Q$, and $\underline{O}$ are obtained, the gain imbalance, quadrature skew, and LO leakage can be calculated with Eqs. 3, 4, and 5.

Equation 7 contains the three unknowns $\underline{G}_I$, $\underline{G}_Q$, and $\underline{O}$, so three independent equations are needed to solve them. This can be accomplished by capturing three samples of the transmitted complex baseband signal, $s(t)$, and the corresponding three samples of the received complex baseband signal, $r(t)$.

The received complex baseband signal, $\underline{r}(t)$, is determined directly by sampling the received signal from a device

under test (DUT) at the output of the I/Q demodulator. However, the transmitted signal, $\underline{s}(t)$, is not readily available. It is synthesized in the measuring device from the received digital data, which can be recovered by demodulating $\underline{r}(t)$. This synthesized signal is denoted as $\underline{s}'(t)$.

Unfortunately, all digital modulation formats are designed so that the absolute phase is always irrelevant. This is necessary because the phase at the receiving end is always random. Hence, there is an infinite number of $s'(t)$ that carry the same digital content as the actually transmitted $\underline{s}(t)$, differing only in phase. They can be rotated into the actually transmitted signal, $\underline{s}(t)$, by a phase shift, δ, shown in Eq. 8.

As a result, the system of equations that is solved is Eq. 9 rather than Eq. 7. The question is whether this phase ambiguity, δ, affects the measurement results or not. Substituting Eq. 8 into Eq. 7 and expanding yields Eq. 10.

Comparing this with Eq. 9 reveals the relationship between $\underline{G'}_I$, $\underline{G'}_Q$, $\underline{O'}$ calculated using the synthesized transmitted signal, $\underline{s}'(t)$, and the actual $\underline{G}_I$, $\underline{G}_Q$, $\underline{O}$ and shown in Eqs. 11-13.

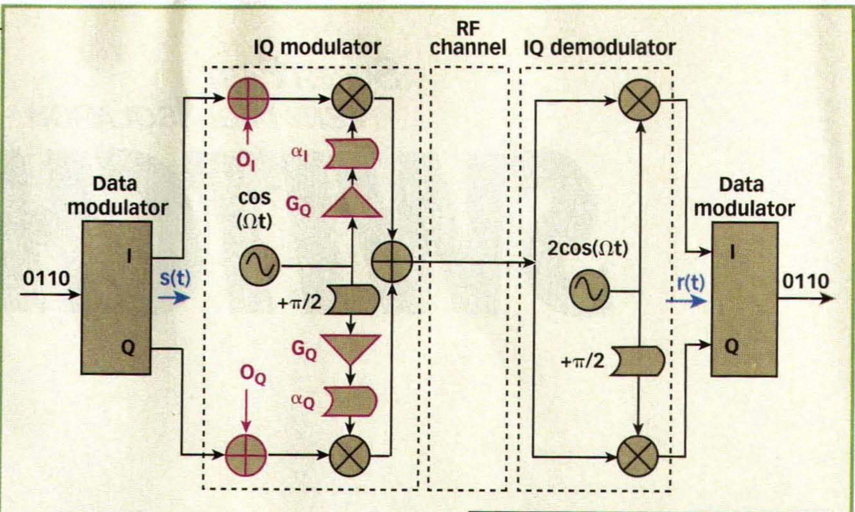
Based on these results, the relationships between the perceived *GainImbalance'*, *QuadratureSkew'*, *LOLeakage'*, and the actual *GainImbalance*, *QuadratureSkew*, *LOLeakage* are shown in Eqs. 14-16.

Equations 14 and 15 show that the calculated gain imbalance and quadrature skew do differ from their actual values if $\delta \neq 0$, that is, if the synthesized transmitted signal, $\underline{s}'(t)$, differs in phase from the actual transmitted signal, $\underline{s}(t)$.

To eliminate this phase ambiguity, the phase reference of the actual transmitted signal, $\underline{s}(t)$, must be entered into the measurement equipment as a reference. Unfortunately, this information is difficult to obtain because it is irrelevant as far as the digital communication is concerned. Therefore, it is rarely standardized and, by the same token, it is not very well documented.

It is possible to determine this phase empirically using an I/Q modulator whose gain imbalance and quadrature skew are some known, but nonzero values. However, different mobile-station models might use different phases. In addition, this phase might change without notice over time.

Worse still, even the assumption that the phase is constant might not be



1. This simple model shows the RF signal chain in a digital wireless communications system, including analog I/Q modulator and demodulator.

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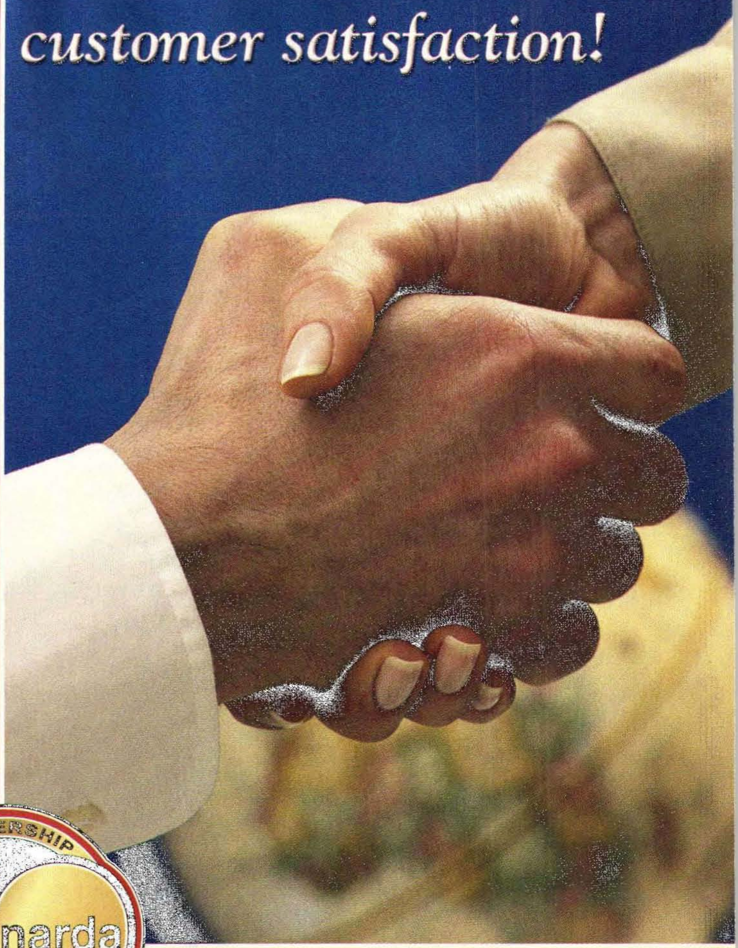

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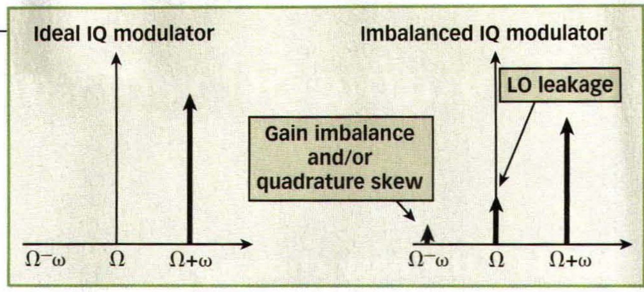
true. It may change randomly or semi-randomly burst by burst. This is the worst-case scenario because it implies that the phase is impossible to track and take into account. As a result, the gain imbalance and quadrature skew figures change randomly burst by burst. This makes I/Q tuning difficult and averaging measurement results (to obtain more accurate results) impossible.

Searching for an alternative measure to replace gain imbalance and quadrature skew, it is helpful to look at how traditional I/Q modulator tuning is performed. The I and Q channels are driven with in-phase and quadrature-phase sine waves. Using complex number notation, this signal is simply a complex single-tone signal at $+\omega$ frequency in Eq. 17.

Using a spectrum analyzer to monitor the output of the I/Q modulator, a perfect I/Q modulator would only show a tone at $\Omega + v$ frequency, where Ω is the LO frequency of the I/Q modulator. An imbalanced I/Q modulator, however, exhibits three tones at the output: the desired component at $\Omega + v$ and two spurious components at Ω and $\Omega - v$ frequencies (Fig. 2). The Ω component is called LO leakage, which, as defined by Eq. 5, represents the ratio of the spurious signal power at Ω to the desired signal power at $\Omega + v$. The $\Omega - v$ component results from both the gain imbalance and the quadrature skew.

Looking at the I/Q imbalance measurements from this perspective, it is interesting that the phase issue is absent. The spectrum analyzer is not synchronized with the signal sources. Each time the spectrum analyzer sweeps through the spectrum; the I/Q modulator is driven by a tone with a different phase, ϕ. Still, a steady consistent reading is apparent when looking at the power of these tones in the signal analyzer. This suggests that a new measure, called "I/Q imbalance," could be used as the ratio of the spurious signal power at $\Omega - \omega$ to the desired signal power at $\Omega + \omega$, analogously to the LO leakage definition. It is easy to obtain its definition by substituting Eq. 17 into Eq. 7, which, after rearrangement, yields Eq. 18.



2. This comparison of ideal and realistic I/Q modulators shows signals resulting from component imperfections.

Notice that this formula describes the received signal at the output of the I/Q demodulator as seen in Fig. 1. The frequency components $\Omega + \omega$, Ω, and $\Omega - \omega$ at the output of the I/Q modulator are mapped to $-\omega$, 0, and $+\omega$ at the output of the I/Q demodulator. With this in mind, the new measure is simply Eq. 19.

Using Eqs. 11 and 12, it can be shown that Eq. 20 results and Eq. 21 follows.

What is needed now is a conversion formula that converts the *GainImbalance'* and *QuadratureSkew'* readings provided by the measurement instruments into meaningful data. This can be obtained from Eq. 21 by using the definitions of gain imbalance and quadrature skew to derive Eq. 22.

One might wonder how this solves the problem of not being able to measure gain imbalance and quadrature skew without knowing the phase reference of the transmitted signal. However, not knowing this phase simply means that the obtained gain imbalance and quadrature skew measures are meaningless if one wants to relate them directly back to the I/Q modulator model of Fig. 1. They can still be used to calculate the I/Q imbalance measure. Since the I/Q imbalance parameter is independent of this phase, these incorrect gain imbalance and quadrature skew parameters yield the same I/Q imbalance measure. So, if the phase of the ideal baseband signal is assumed incorrectly, or worse still, if this phase changes randomly, the I/Q imbalance figure still remains a constant value.

The I/Q imbalance measure defined by Eq. 22 is completely analogous to the image-frequency power measure traditionally used for tuning I/Q modulators. This means that the same tuning procedures developed for this traditional technique can be used with the new I/Q imbalance metric.

The I/Q imbalance metric has the great advantage that it makes no assumption about the test signals driving the I/Q modulator. It can be a sine wave, but equally, it can be a complex modulated signal. It is therefore possible to perform the I/Q modulator tuning procedure while transmitting real data. This avoids the need for a special test mode required by the traditional method. MRF

$$QuadratureSkew' = \angle\left(\frac{10^{\frac{Gain\,\mathrm{Im}\,balance}{20}} e^{jQuadratureSkew}\tan\delta}{1 + j10^{\frac{Gain\,\mathrm{Im}\,balance}{20}} e^{jQuadratureSkew}\tan\delta}\right) \quad (15)$$

$$LOLeakage' = LOLeakage \quad (16)$$

$$\underline{s}(t) = A\left(\cos(\omega t + \phi) + j\sin(\omega t + \phi)\right) = Ae^{j\omega t + \phi} \quad (17)$$

$$\underline{r}(t) = \frac{\underline{G}_I + \underline{G}_Q}{2} Ae^{j\omega t + \phi} + \frac{\underline{G}_I - \underline{G}_Q}{2} Ae^{-(j\omega + \phi)} + \underline{O} \quad (18)$$

$$IQ\,\mathrm{Im}\,balance = 20\log_{10}\left|\frac{\underline{G}_I - \underline{G}_Q}{\underline{G}_I + \underline{G}_Q}\right| \quad (19) \qquad \left|\frac{\underline{G}'_I - \underline{G}'_Q}{\underline{G}'_I + \underline{G}'_Q}\right| = \left|\frac{\underline{G}_I - \underline{G}_Q}{\underline{G}_I + \underline{G}_Q}\right| \quad (20)$$

$$IQ\,\mathrm{Im}\,balance = 20\log_{10}\left|\frac{\underline{G}'_I - \underline{G}'_Q}{\underline{G}'_I + \underline{G}'_Q}\right| \quad (21)$$

$$IQ\,\mathrm{Im}\,balance = 20\log_{10}\left|\frac{1 - 10^{\frac{Gain\,\mathrm{Im}\,pedance'}{20}} e^{jQuadratureSkew'}}{1 + 10^{\frac{Gain\,\mathrm{Im}\,balance'}{20}} e^{jQuadratureSkew'}}\right| \quad (22)$$

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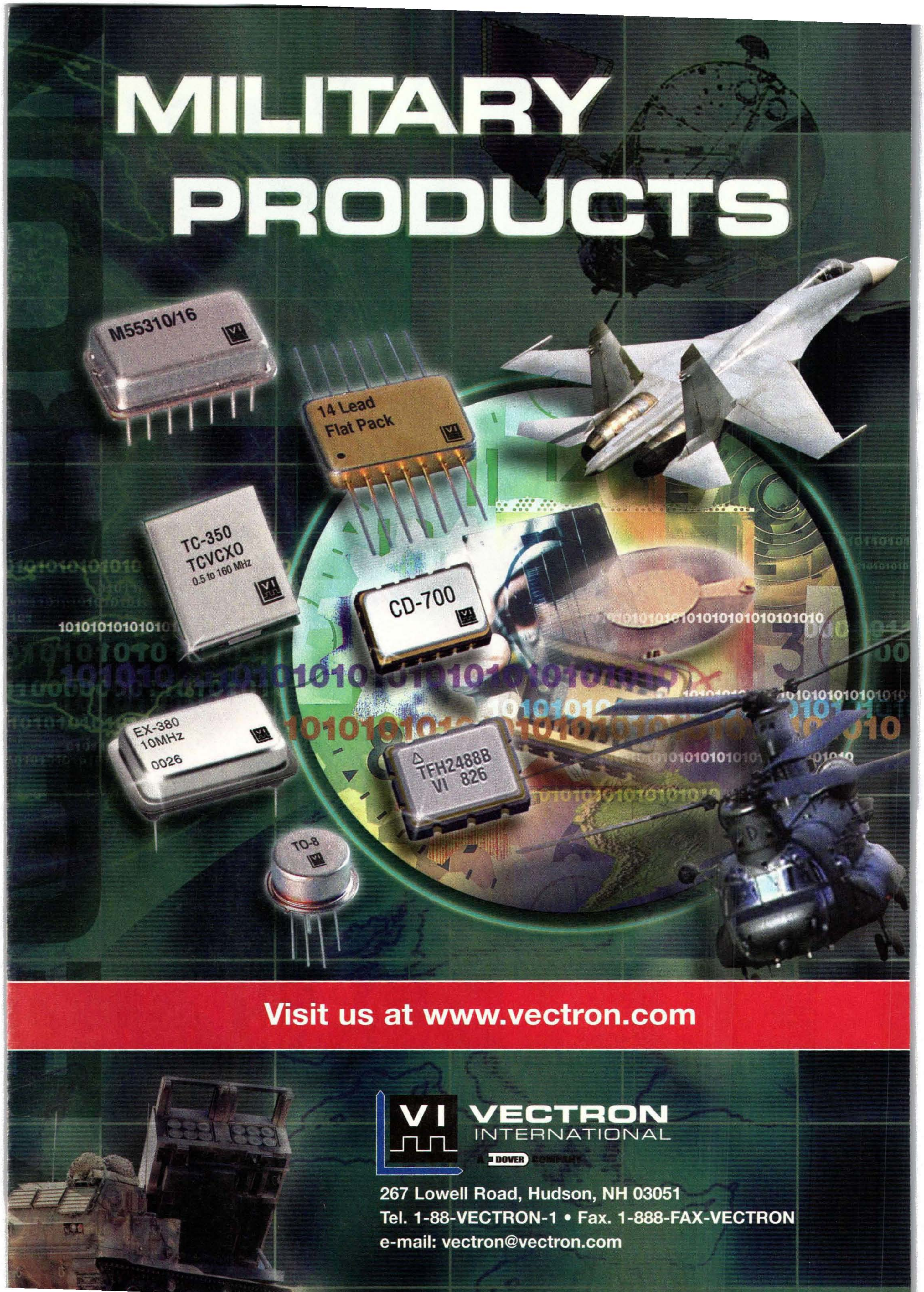
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Design Considerations For Microwave RF Repeaters

Microwave RF repeaters are designed to transfer signals from one radio terminal to another without loss of quality, data, or traffic, while compensating for multipath and fading loss.

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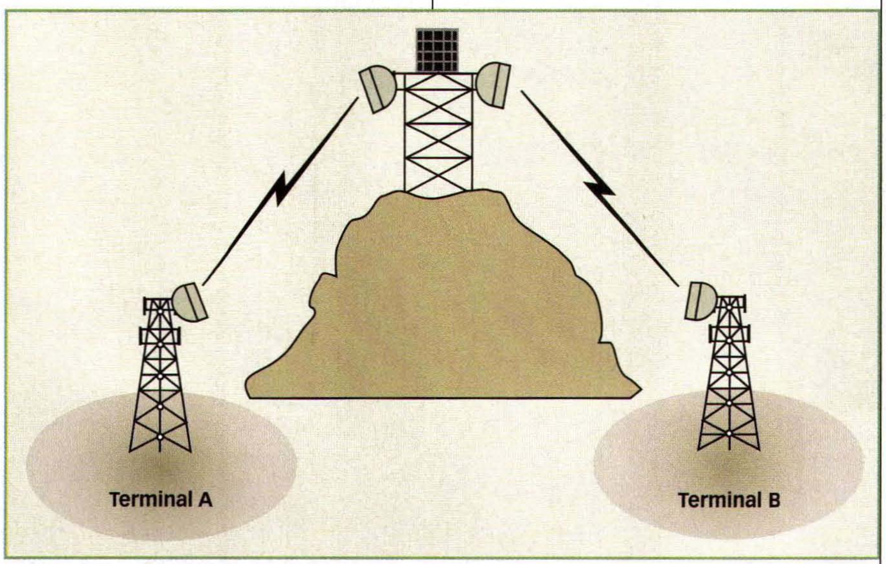
Microwave on-frequency RF repeaters are commonly used by telecommunications system operators to reliably and cost-effectively relay radio signals at remote locations, typically mountaintops and when bypassing obstructed paths. Understanding the use of microwave on-frequency repeaters requires an understanding of some basic operating concepts and how to apply the latest techniques.

The recently updated RF Repeater Applications Design Tool from Peninsula Engineering Solutions (San Ramon, CA) is a useful program for understanding the operation and application of microwave repeaters. In their simplest form, microwave RF repeaters are fairly simple, linear, on-frequency gain blocks. They can support a wide range of modulation formats and traffic capacity, and the use of channel filters can set the required bandwidth while supporting standard frequency plans. The repeaters, which are often powered by solar- or wind-based energy sources, receive and retransmit signals without loss in quality or capacity.



1. A microwave repeater link is designed to transfer signals from one terminal station to another without loss of traffic or signal performance.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Organizations that use microwave repeaters include telephone companies, wireless operators, energy companies (water, gas, electric), government agencies (including national, state, county, and local agencies), military, aviation, and national security organizations. Such users expect reliable operation; areas of prime concern include path reliability, repeater-equipment reliability, and power-equipment reliability.

Path reliability normally has to meet the same standards as the rest of the microwave radio relay system. Reliability objectives are often stated on a per hop basis or end-to-end. The most often-used reference objective is the AT&T Short Haul standard, which is defined as 99.98 percent or 6400 s per 250-mile section, end-to end, two-way, with path fading and equipment annual outage combined. Path fading is normally allocated one-half the annual outage budget, 99.99 percent or 3200 s per 250-mile section. The objective applied to each hop is apportioned on a distance ratio basis: d/250 mi. For example, a 30-mile path would have a two-way outage objective of 384 s or less. Some organizations may require more stringent path reliability objectives, such as 99.9999 percent per hop in heavy route applications.

Fading mechanisms considered include fading due to multipath phenomena, obstructions, and rain attenuation. Equipment and power-source reliability demands are dealt with through a combination of highly reliable components and modules plus designs that incorporate redundancy and protection. For example, Peninsula Engineering Solutions addresses these considerations with protected, soft-fail amplifiers and dual, redundant electric power systems as a minimum approach. Supervisory alarm equipment provides reporting of failures or degraded conditions often with enough early warning time for corrective actions to be taken.

The path-transmission-reliability models used for RF repeaters are the same as for most terrestrial, line-of-sight microwave paths. The classic model is the Vigants-Barnett model, with improvements by W. Rummler and others. The International Telecommunications Union (ITU) ITU-R models are frequently used outside of North America. Rain attenuation is normally considered above 9 GHz. Both the Crane and ITU-R rain models can be applied for estimating the path loss due to rain attenuation.

2. This photograph shows a typical microwave repeater station.

Some of the assumptions that can be applied to microwave RF repeater models include the idea that hops fade independently, so each hop can be calculated separately. Also, rain outages affect two-way communications, and multipath outages do not occur during rainfall. In addition, space- and frequency-diversity techniques can be applied for improved performance in one or two directions.

Determining the equivalent receiver threshold value for a microwave RF repeater is one of the more demanding differences compared to standard transmission engineering. Since microwave RF repeaters do not demodulate traffic, only amplify it, they do not have a designated threshold value even for a specific modulation and traffic capacity. Rather, the equivalent receiver threshold is relative to the terminal radio's threshold and associated noise figure: the net path loss plus the repeater's noise figure and maximum gain. The approach has been to use the cascaded noise figure equation as the basis for determining the equivalent repeater threshold or "minimum receive power":

SEE EQUATION IN BOX BELOW

where:

$Gain_R$ = the RF repeater maximum gain (in dB);

NF_R = the RF repeater noise figure (in dB);

NF_T = the terminal radio noise figure (in dB);

NPL = the unfaded net path loss

$$10\times\log\left[10^{\left(\frac{NF_R}{10}\right)}+\frac{10^{\frac{NF_T+NPL+PAD_{Out}}{10}}-1}{\left(10^{\frac{Gain_R}{10}}\right)}\right]+Min\_Rx\_Pwr_T-NF_T+PAD_{in}$$

between the RF repeater transmitter and terminal radio receiver (in dB);

$Min\_Rx\_Pwr_T$ = the terminal radio threshold (in dBm)

PAD_{In} = the RF repeater input attenuator pad attenuation (in dB); and

PAD_{Out} = the RF repeater output attenuator pad attenuation (in dB).

The RF repeater receive flat fade margin (FFM, dB) thus becomes:

FFM, dB = [nominal receive signal level (dBm) – minimum receive power (dBm)]

Linear RF repeaters are compatible with a wide range of modulation formats. The transmit power level for a particular repeater model depends on two parameters: the frequency modulation (FM) or fully rated power level and the backoff amount for the modulation used by the terminal radios. The **table** offers examples of transmit power setting per modulation type. The appropriate transmit power level is selected by looking up the modulation format. Terminal radio traffic capacity is not a consideration when selecting the RF repeater transmit power level.

Frequently, the RF repeater transmit power rating will be less than the associated terminal radio. This difference is one cause of asymmetrical receive signal levels and fade margins per hop. Since the fade margins are different per direction per hop, it is necessary to calculate the reliability per direction per hop as well.

Normal transmission engineering practice is to assume that each path fades independently. This makes sense when dealing with radio terminals that have high-enough gain and automatic gain control (AGC) to compensate for 60 dB unfaded net path loss plus 40 to 50 dB deep fast fades and maintain full transmit power. Microwave repeaters normally have less gain and AGC as a consequence of on-frequency operation. Typical microwave radio terminals have 100 to 120 dB system gain and 50 dB down-fade AGC range, while microwave RF repeaters have 50 to 70 dB system gain and 20 dB down-fade AGC range.

Comparing modulation formats

MODULATION	BACKOFF	RATED POWER LEVEL
Fully rated, FM FSK, MSK	0 dB	28 dB
4PSK, QPSK, OQPSK	2	26
16QAM	6	22
64QAM	10	18
128QAM	12	16
32TCM	9	19
128TCM	11	17

A classic characteristic of multipath fading is that the deeper the fade, the shorter the fade duration below a given depth. This is also known as "time below level." Fades greater/deeper than 30 dB are very short and hence less likely to occur simultaneously. Shallow fades less than 10 dB occur often with some paths constantly in shallow fade. When considering shallow fades, it is somewhat likely that multiple hops will experience simultaneous fades less than 10 dB.

A microwave RF repeater with 10 to 20 dB down-fade AGC reserve will fully compensate for this shallow fading. Path fading in excess of the RF repeater's AGC reserve will be passed on to the following terminal radio. When more than one microwave repeater is used in tandem, the cumulative shallow fading may exceed the cumulative AGC reserve. As an example, consider three microwave repeaters in tandem on a four-hop section between terminals. Each repeater has 5 dB of AGC reserve and each hop has 8 dB of shallow fading. The end-to-end link will fade $(4 \times 8\text{ dB}) - (3 \times 5\text{ dB}) = 32 - 15 = 17$ dB. While the terminal radios can easily compensate for 17 dB of fading, the concept does affect designs that use multiple tandem microwave repeaters. A conservative approach is to limit the number of tandem microwave repeaters to three or possibly four when the expected fading conditions are favorable.

Diversity improvement techniques apply to microwave RF repeaters much in the manner in which they are used with standard microwave radios. It is important to consider that most microwave RF repeaters do not directly provide diversity switching or combining. Terminal microwave radios associated with microwave repeaters provide diversity switching or combining for the whole multihop microwave repeater section. Microwave RF repeaters supporting diversity-receive functions must provide two or more orthogonal channels to the terminal radio receivers. Finally, space diversity can be applied in one direction.

Frequency diversity is somewhat simpler to understand than space diversity. In frequency diversity, two channels on different frequencies carry the same traffic information. Two receivers switch between or combine the two channels to provide the best-quality signal to the receiver demodulators. A microwave RF repeater supporting frequency diversity only needs two parallel amplifying channels per direction, one on each frequency. The two frequencies are the two orthogonal channels. Frequency-selective or multipath fading on any hop in the microwave repeater section is passed to the two receivers at the terminal radio where the diversity action takes place. Diversity receivers on microwave repeater sections work normally except they may be "busier" than normal due to dealing with fades from several hops. Microwave repeaters supporting frequency diversity can be used in tandem without undue concern.

Space diversity is used in several ways with a microwave RF repeater. The first is simple receive diversity at the terminal microwave radio. Here, space diversity can be implemented as a "one-way" improvement in the path from the microwave repeater to the terminal. What about the other direction? It should be remembered that a microwave repeater often has less transmit power than the terminal radio. The resulting asymmetrical fade margins can be balanced better by provisioning space diversity receive at the terminal radio sites especially on the longer hops.

The second application of space

diversity with RF repeaters is in the case where improvement is needed in both directions. Space diversity receive at the microwave repeater works best under the following conditions: the path configuration includes one RF repeater; the paths are one long hop needing improvement and one short hop having lower fading probability; and normal space-diversity antennas are provisioned on the longer path. RF repeaters configured for space diversity may use either a co frequency, dual-polarized pair of links on the short path or in the case of hybrid space- and frequency-diversity techniques, two frequency channels. Either approach provides the necessary orthogonal channels. There may be additional considerations for the co-frequency, dual-polarized configuration regarding antenna cross polarization discrimination (XPD) and receiver co-channel carrier-to-interference (C/I) or threshold-to-interference

[*A classic characteristic of multipath fading is that the deeper the fade, the shorter the fade duration below a given depth.*]

(T/I) values. Hybrid space and frequency diversity can be easily applied in frequency-diversity routes that need extra improvement. The resulting hybrid improvement is the product of the two individual improvement factors.

On-frequency microwave repeaters must manage the feedback or echo signals that can occur on site. Management is mainly focused on the antennas selected and their relative mounting locations. Digital radios typically require a co-channel C/I or T/I value of 25 to 40 dB depending on modulation, traffic capacity, forward-error correction (FEC), and receiver equalization. The antenna isolation C/E should be greater than the radio co-channel C/I or T/I specification. Antennas with adequate front-to-back (F/B) ratios and side lobe suppression are selected for each application. The task of the transmission engineer is often to select the most economical antenna configurations for the project. Often one or more antennas may be size constrained as well. The design tool used by Peninsula Engineering Solutions uses a proprietary C/E prediction method that results in recommending antennas that meet the needs of the project with enough margin to work in the real-world field installation. The recommended antennas are not "over engineered."

Antenna spacing is an available technique that doesn't add much or any cost to projects. When the repeater site antennas are separated by greater dis-

tances, either horizontally or vertically, the antenna isolation increases. Towers with 8-to-12-ft faces are often provisioned at RF repeater sites to provide greater antenna separation.

Links using microwave RF repeaters typically require antennas one size larger than if the repeater site used back-to-back microwave radios. Considering the economics of the RF repeater site with remote alternative power sources, the increased antenna costs are easily accommodated.

A significant consideration to selecting microwave RF repeaters is their low power consumption and optimization for alternative energy sources. Duplex microwave RF repeaters may only use 25 to 50 W of DC power. Reasonably sized solar arrays and storage batteries can provide reliable site power almost anywhere in the world. Solar electric battery systems are designed for the particular site location and microwave repeater load. Most systems have a battery reserve time of 7 to 10 days without solar charging. Adding a 500-W wind turbine generator can provide that extra confidence that battery-charging power is being generated during prolonged storms. Currently available solar panels have life expectancies of 25 years or more. Solar rated batteries typically last 5 to 8 years due to the daily charge and discharge cycles. Charge controller and wind turbines normally last 20 years or more at remote sites.

Beyond the typical duplex configurations where one frequency is used per direction, RF repeaters may be configured for multiline service such as 2 + 1 to 7 + 1 or in a Y-junction function where one repeater site may connect three end points.

Certain higher capacity radios have spectral bandwidths that fill the available channels. When the radio channel and the repeater filter bandwidth are close, it may be beneficial to provision delay equalizers in the RF repeater. The delay equalizers can compensate for the parabolic group delay shape of the channel filters that otherwise may cause distortion and a low error rate.

Microwave RF repeaters follow the same basic design rules as standard microwave radios with a few specific techniques needed for successful applications. Microwave repeater design tools can supplement commercial programs in automating the transmission engineering work. Microwave RF repeaters can meet or exceed the stringent system reliability demands of the user organizations. For more information on repeaters, or a free copy of the RF repeater applications design tool, contact the author or Peninsula Engineering Solutions. MRF

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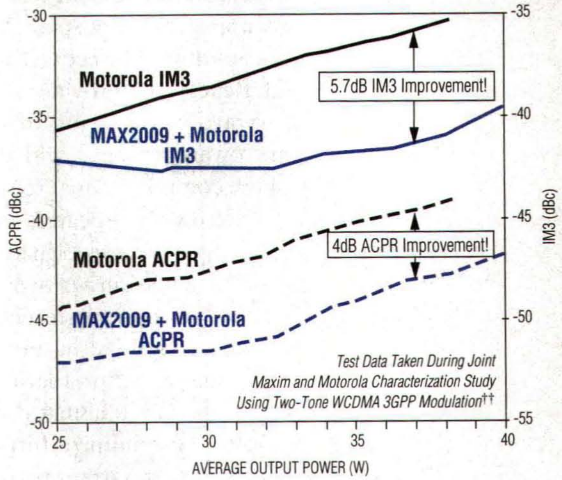

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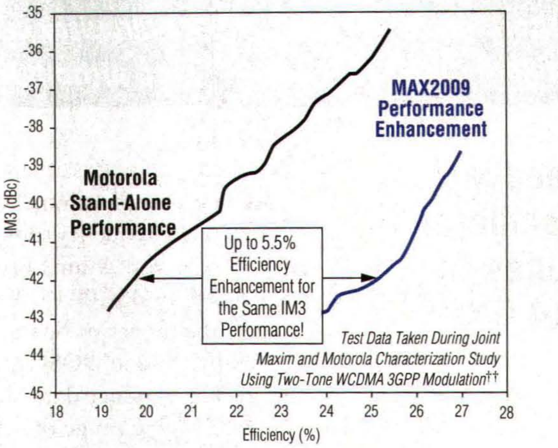

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Scrutinize Selective Fading In Microwave Repeaters

SELECTIVE FADING or dispersive fading is one of the phenomena that affect the performance of a microwave radio relay link. Selective fading is caused by multipath effects with a time delay short enough to cause signal cancellation in only part of a wideband channel. The signal cancellation can result in signal waveform distortion, an increase in the system bit-error rate (BER), and an ultimate loss of transmitted data. An application note from Peninsula Engineering Solutions "Selective Fading in Microwave RF Repeaters," provides a simple and straightforward explanation of selective fading in microwave repeaters, with very little promotion of the company's own repeater product line.

Microwave repeaters are designed to linearly amplify received signals and retransmit them to the next point in a network. Since the repeater normally boosts any received signal within its designed bandwidth, any selective-fade distorted signal within that bandwidth will also be amplified and transmitted along with the desired signals. To minimize further loss of performance, it is important that the power amplifier remain linear under all signal conditions. When nonlinearities in the repeater are combined with distorted signals, more complex distortion, including amplitude-modulation/pulse modulation (AM/PM) distortion and AM/PM with delay (AM/PM + delay) can result. While some of the effects of these distortions can be corrected at a terminal receiver, it may not always be possible to recover lost data due to the delerterious effects of the distortion.

Automatic-gain-control (AGC) circuitry (employed in the company's line of microwave repeaters) can compensate for path fading by dynamically controlling the output-power level, although the different types of level detectors used with these circuits can provide different results and must be carefully considered.

Copies of the four-page application note, which includes a comprehensive listing of reference literature on radio engineering and microwave fading, are available for free download from the company's website.

Peninsula Engineering Solutions, 39 Grand Canyon Lane, San Ramon, CA 94583; (925) 901-0103, FAX: (925) 901-0403, Internet: www.peninsulaengineering.com.

Automatic-gain-control (AGC) circuitry can compensate for path fading by dynamically controlling the output-power level, although detector types must be carefully considered.

Feedback Oscillator Tunes To 8.4 GHz

CLASSIC FEEDBACK OSCILLATORS can provide reasonably low phase noise while covering a relatively wide frequency tuning range. An application note from Mimix Broadband, Inc., simply titled "9OSC0315 Application Note," details the theory of operation and use of the company's model 9OSC0315 feedback oscillator. The product data sheet for the device shows a tuning range of 5.5 to 8.4 GHz with specified output power of +4 dBm. The specified single-sideband (SSB) phase noise at an offset frequency of 10 kHz is –56 dBc/Hz while the phase noise at an offset of 100 kHz improves to –80 dBc/Hz.

As the note explains, the frequency at which a basic feedback oscillator will oscillate is the frequency for which the total phase shift in the feedback loop, introduced through the oscillator's active device (amplifier) and transmission lines, is some multiple of 2π. The oscillation at this frequency will be maintained if the magnitude of the feedback loop gain is greater than or equal to unity. The loop gain is usually greater than unity and the amplitude of the oscillator is then limited by the saturation of the active device or amplifier.

The operating frequency of a feedback amplifier is determined by the total phase delay introduced in the amplifier and the feedback path. Since it is difficult to know the exact delay in a circuit, especially at microwave frequencies, it is also difficult to accurately predict the oscillation frequency of a feedback oscillator. Delays through the amplifier can vary significantly from device to device, and even bond wires in a monolithic-microwave-integrated-circuit (MMIC) oscillator such as the model 9OSC0315 can significantly influence the final frequency.

Copies of the nine-page application note, which includes details on phase tuning and the addition of divider circuitry, are free upon request from the company.

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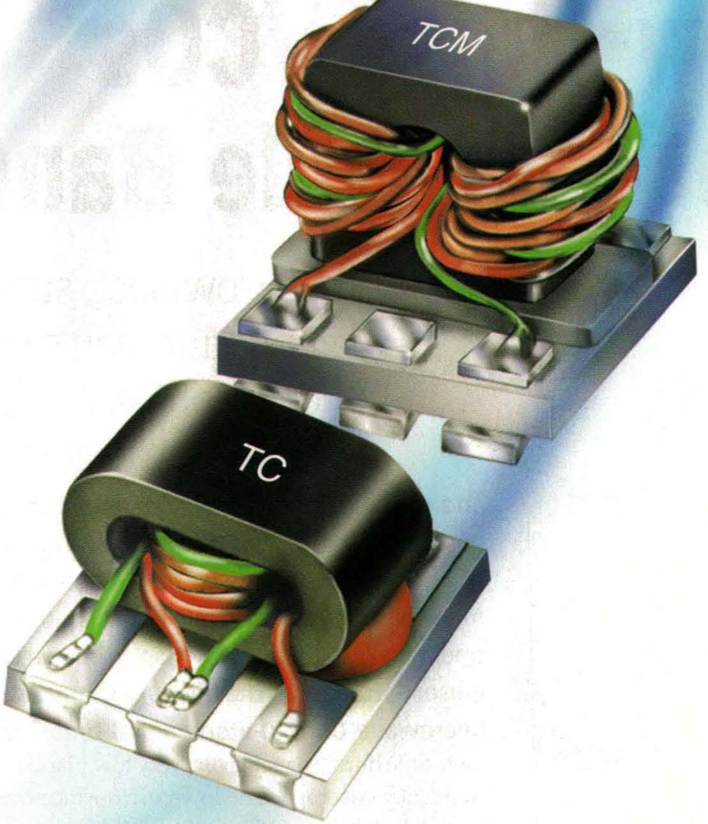

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TC1-1-13M	1G	4.5-3000	4.5-1000	.99
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TC3-1T	3A	5-300	5-300	1.29
TC4-1T	4A	.5-300	1.5-100	1.19
TC4-1W	4A	3-800	10-100	1.19
TC4-14	4A	200-1400	800-1100	1.29
TC8-1	8A	2-500	10-100	1.19
TC9-1	9A	2-200	5-40	1.29
TC16-1T	16A	20-300	50-150	1.59
TC4-11	50/12.5D	2-1100	5-700	1.59
TC9-1-75	75/8D	0.3-475	0.9-370	1.59

LEADS Plastic Base (actual size)

MODEL	Ω Ratio & Config.	Freq. (MHz)	Ins. Loss❖ 1dB (MHz)	Price $ea. (qty. 100)
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TCML1-11	1G	600-1100	700-1000	1.09
TCML1-19	1G	800-1900	900-1400	1.09
TCM2-1T	2A	3-300	3-300	1.09
TCM3-1T	3A	2-500	5-300	1.09
TTCM4-4	4B	0.5-400	5-100	1.29
TCM4-1W	4A	3-800	10-100	.99
TCM4-6T	4A	1.5-600	3-350	1.19
TCM4-14	4A	200-1400	800-1000	1.09
TCM4-19	4H	10-1900	30-700	1.09
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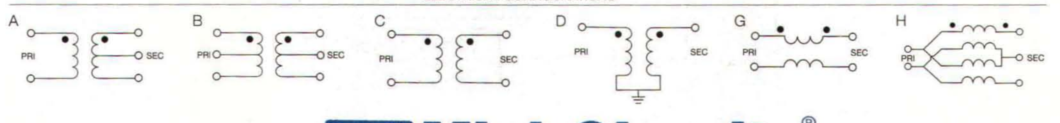

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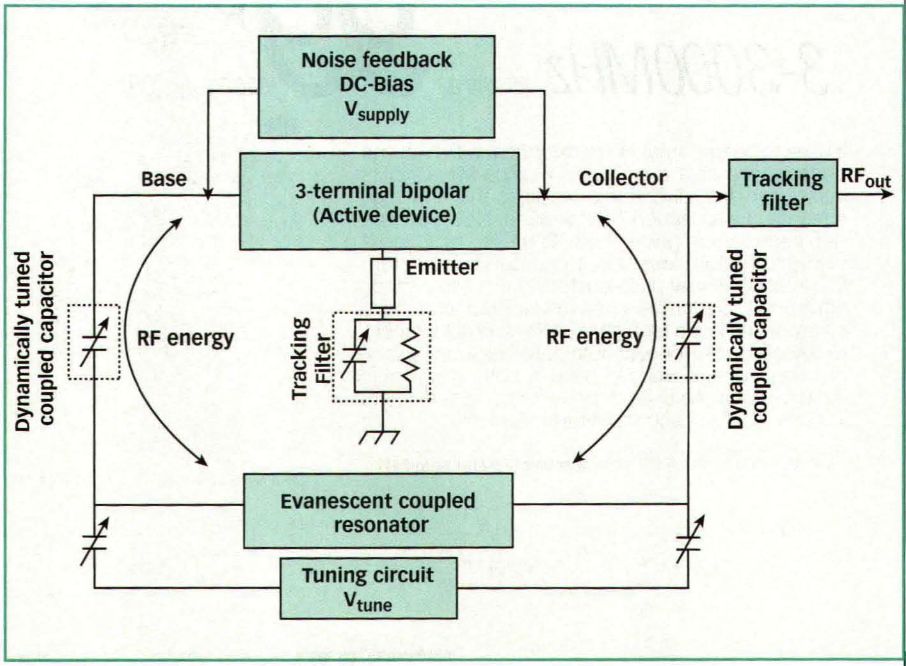
cover story

Low-Noise VCOs Conquer Wide Bands

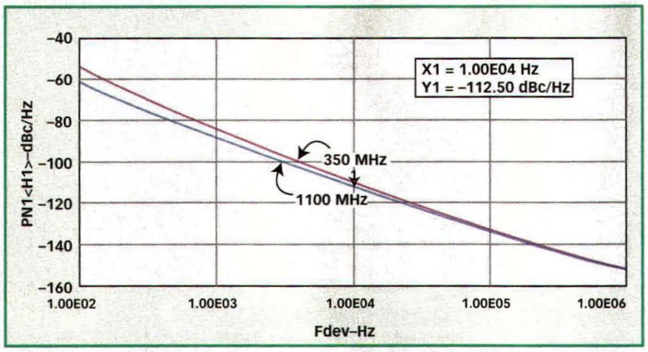
These low-cost, surface-mount sources offer better than octave tuning ranges and low phase noise while consuming minimal power through 4200 MHz.

ULRICH L. ROHDE
K. JUERGEN SCHOEPF
AJAY KUMAR PODDAR
Synergy Microwave Corp., 201 McLean Blvd., Paterson, NJ 07504; (973) 881-8800, FAX: (973) 881-8361, e-mail: sales@synergymwave.com, Internet: www.synergymwave.com.

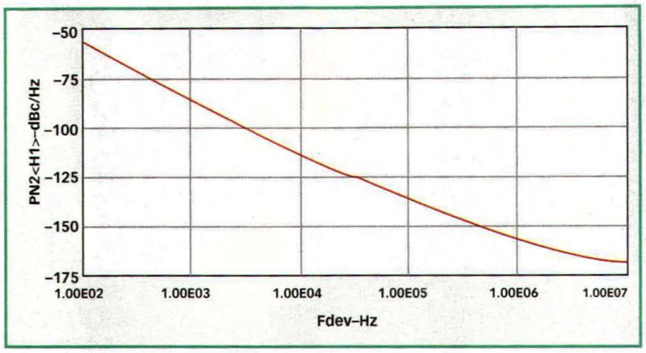
bandwidth and phase noise are two of the leading requirements for signal-generating components in modern communications equipment. Voltage-controlled oscillators (VCOs) have been the frequency source of choice for many wired-, wireless-, and optical-communications systems, even though traditionally limited to less than an octave tuning range to maintain low phase noise. Fortunately, the new DCFO and DCMO series of VCOs from Synergy Microwave Corp. (Paterson, NJ) break with tradition and overcome the long-time hurdle of achieving very low phase noise while also delivering broadband frequency coverage. The low-cost, surface-mountable VCOs are currently available in bands from 350 to 4200 MHz.



1. This block diagram shows the discrete-device approach with evanescent-coupled resonator used in the DCFO and DCMO broadband VCOs.



2. This phase-noise plot shows the performance of a DCFO oscillator for two carrier frequencies at the edges of the full bandwidth with bias of 28 mA at +5 VDC.

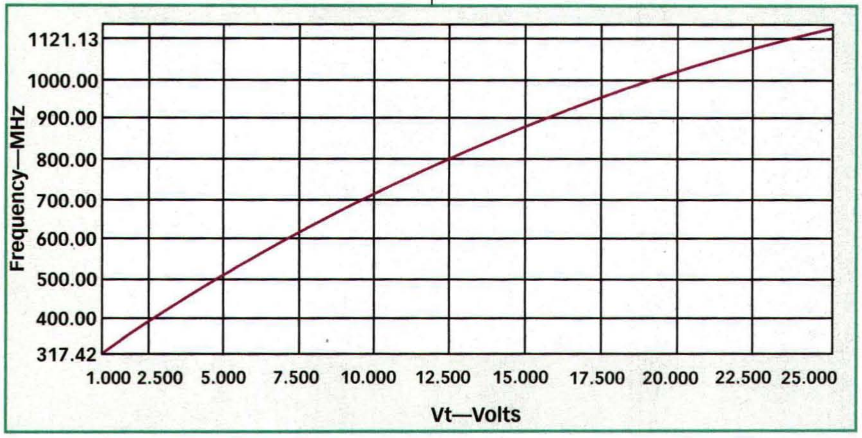


3. This phase-noise plot shows how the performance of a DCFO oscillator improves with a +12-VDC (28-mA) supply.

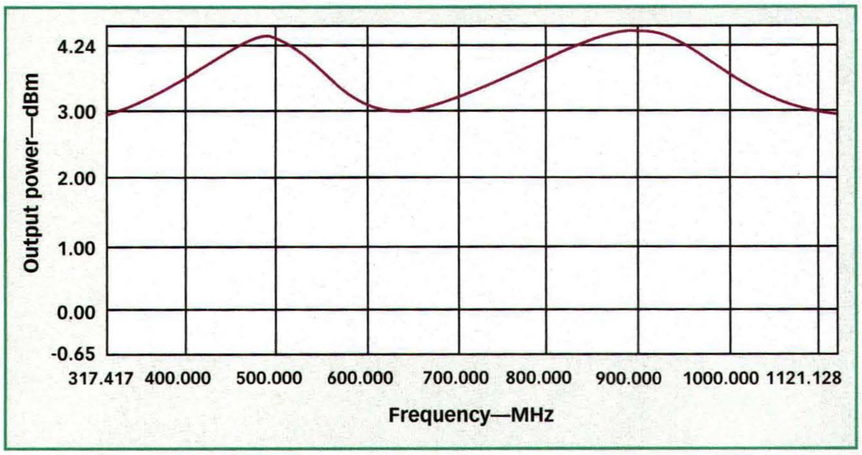
The new VCOs are particularly well suited for the new families of cellular handsets and base stations with Universal Mobile Telephone Systems (UMTS) requirements. Increased-bandwidth coverage is needed in support of combined voice, data, and wireless Internet services, yet the frequency source must also deliver very low levels of single-sideband (SSB) phase noise in order to reliably handle the complex digital modulation employed in these systems while also stably operating within tightly spaced communications channels. In a digital wireless-communications system, excessive phase noise can cause degradation in the effective system bit-error rate (BER), resulting in a loss of transmitted/received data and a loss of voice and data performance as perceived by the wireless customer.

Wideband tuning and low phase noise have long been assumed as opposing design targets. A decrease in VCO phase noise generally meant a decrease in tuning bandwidth, due to the problem of simultaneously controlling the loop parameters and optimizing the time average loaded quality factor (Q) of the VCO resonator over the tuning range. The tuning range of the oscillator generally influences the phase noise and typically there is a trade-off between the continuous tuning range of VCOs and the amount of phase noise generated by the varactor capacitance modulation.[1] On the other hand, the requirements for low-noise performance over a broad (more than an octave) frequency range are typically demanding. Thus, there exists a need for method and circuitry for improving the phase-noise performance over a wide tuning frequency range, typically more than an octave-band tuning range.

Although a great deal of progress has been made in recent years in monolithic, integrated-circuit (IC) VCOs, with a desire to fabricate completely integrated radio front-end circuitry for large-volume communications applications (such as cellular handsets), the best performance levels are still the domain of discrete-device VCOs. In



4. This plot shows the extremely linear tuning response of a DCFO oscillator with tuning voltages from 1 to 25 V.



5. Although specified for +1 dBm and ±3.5 dB flatness, the output power of a typical DCFO oscillator is much higher and flatter over frequency.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addition to high performance, discrete VCOs offer advantages such as superior performance, tremendous design flexibility and versatility, faster time-to-market, low cost, and reduced risk.

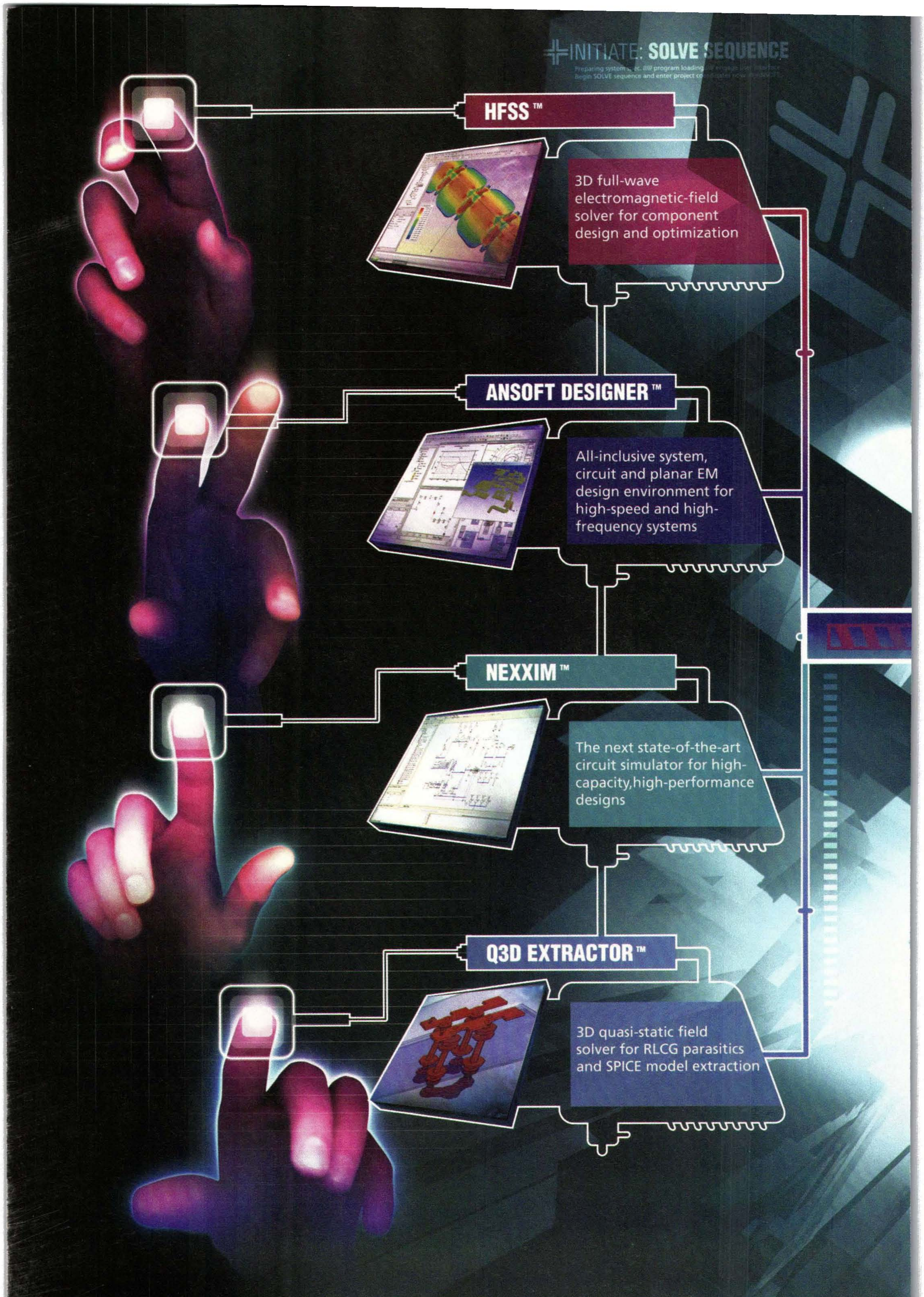
The discrete-device approach was used in the development of the DCFO and DCMO series oscillators. The sources employ a novel oscillator topology (for which a patent has been applied) based on an evanescent-mode dynamic coupled resonator **(Fig. 1)**.[2] The design approach has resulted in wideband VCOs capable of delivering stable, low-noise output signals **(Figs. 2 and 3)** over temperature ranges as wide as –40 to +85°C with extremely linear tuning response **(Fig. 4)**. The **table** offers a brief overview of some of the new VCOs. As the table shows, the new VCOs offer tuning ranges as wide as 2400 MHz, but without sacrificing phase-noise performance. Using a dynamic tracking filter at the output, harmonics can be suppressed by better than –30 dBc.

The wideband VCOs at a glance

SERIES/SIZE	FREQUENCY RANGE (MHz)	TYPICAL PHASE NOISE OFFSET 10 kHz FROM CARRIER	TYPICAL PHASE NOISE OFFSET 100 kHz FROM CARRIER
DCFO	350 to 1100	-112 dBc/Hz	-132dBc/Hz
DCMO	500 to 1700	-99 dBc/Hz	-120dBc/Hz
DCMO	1500 to 3500	-92 dBc/Hz	-112 dBc/Hz
DCMODCFO	1500 to 3500	-90 dBc/Hz	-110 dBc/Hz

As an example of the DCFO series, model DCFO-35105 accepts tuning voltages from 0 to +25 VDC to cover a total range of 350 to 1050 MHz (700 MHz). The bias requirements are no more than 35 mA at +5 VDC. The tuning sensitivity is typically 20 to 48 MHz/V. With output power of +1 dBm **(Fig. 5)**, the VCO exhibits typical phase noise of –112 dBc/Hz offset 10 kHz from the carrier and –132 dBc/Hz offset 100 kHz from the carrier. Harmonic suppression for this model is specified at –10 dBc, although typical performance is much better **(Fig. 6)**. Maximum frequency pulling is 4 MHz for a 1.75:1 VSWR load while maximum frequency pushing is 2 MHz/V. The VCO is supplied in a surface-mount package with slotted metal cover measuring just 0.91 × 0.91 × 0.305 in.

As an example of the higher-frequency DCMO series VCOs, the model DCMO-190410 covers a tuning range of 1900 to 4100 MHz by means of tuning voltages from 0.5 to 20.0 V. The typical bias requirements are 35 mA (maximum) and +5 VDC, and the typical tuning sensitivity is 100 to 200 MHz/V. With minimum output power of +3 dBm across this wide tuning range, the VCO delivers typical phase noise of –90 dBc/Hz offset 10 kHz from the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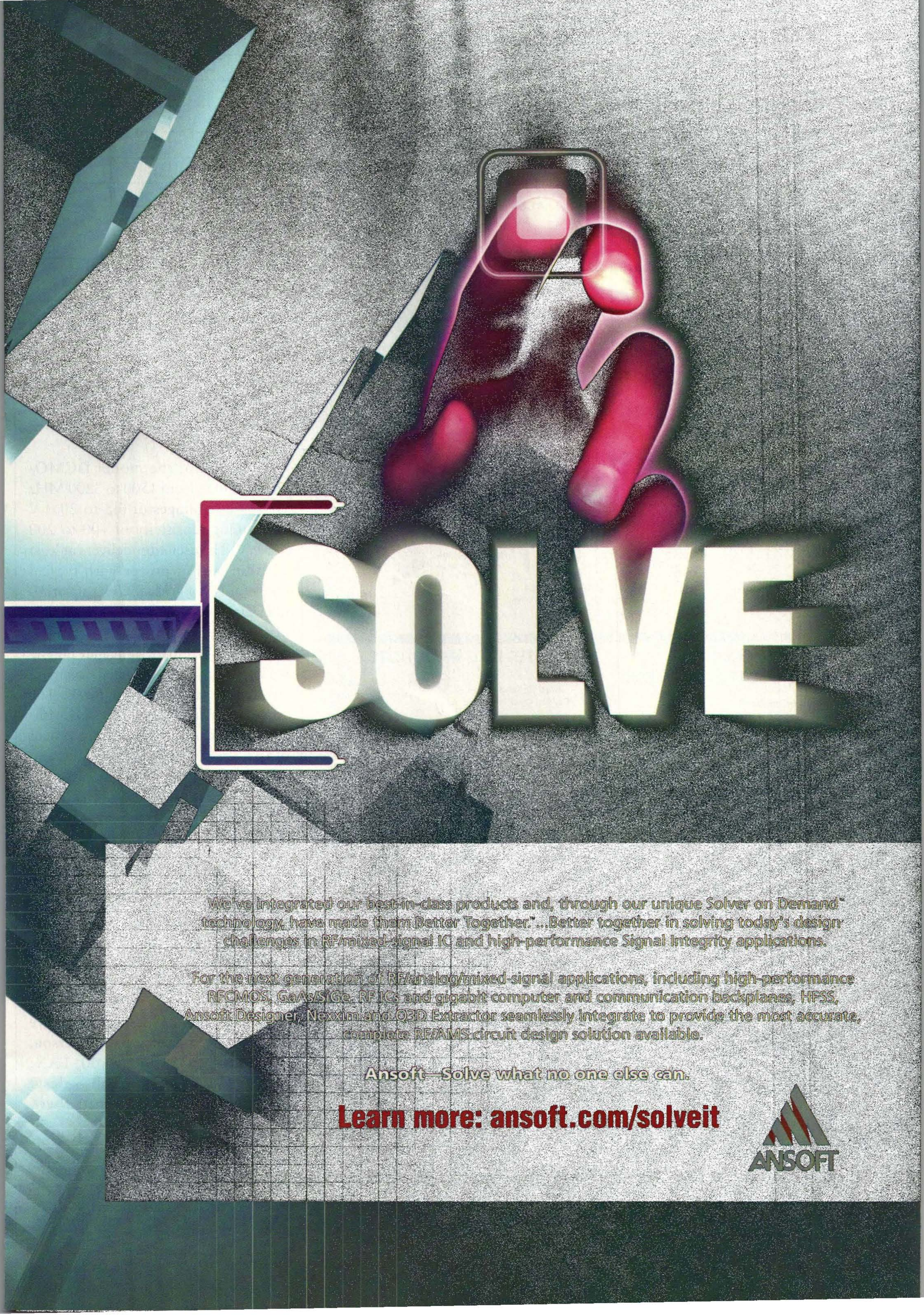
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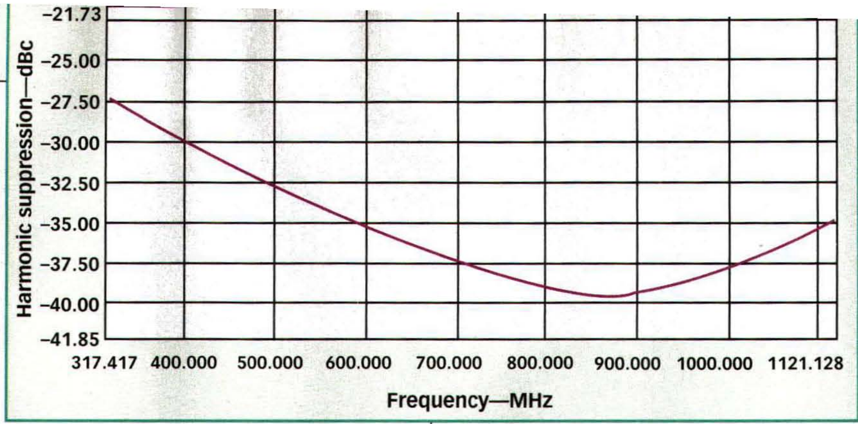
carrier and –110 dBc/Hz offset 100 kHz from the carrier. Maximum frequency pulling is 14 MHz for a 1.75:1 VSWR load while maximum frequency pushing is 7 MHz/V. The VCO is supplied in a surface-mount package with slotted metal cover measuring just 0.50



6. Harmonic suppression in a DCFO oscillator is specified as –10 dBc, although measured performance clearly exceeds the specification.

× 0.50 × 0.25 in. Both the DCFO-35015 and the DCMO-190410 have an operating temperature range of –30 to +75°C.

In between, the model DCMO-150320 tunes from 1500 to 3200 MHz via tuning voltages of 0.5 to 20.0 V and tuning sensitivity of 100 to 200 MHz/V. The oscillator delivers at least –2 dBm output power with typical phase noise of –92 dBc/Hz offset 10 kHz from the carrier and –112 dBc/Hz offset 100 kHz from the carrier. With a 1.75:1 VSWR load, frequency pulling is a maximum of 12 MHz; the maximum frequency pushing is 6 MHz/V. The company also offers model DCMO-60170 with minimum output power of +3 dBm from 600 to 1700 MHz and phase noise of –99 dBc/Hz offset 10 kHz from the carrier and –120 dBc/Hz offset 100 kHz from the carrier.

In short, these VCOs combine the much desired wide tuning ranges required for multimode operation in next-generation cellular-communications handsets and infrastructure equipment with the low phase noise of much narrower-band sources. The combination should allow designers to make use of a single VCO where two or more were used in the past. **Synergy Microwave Corp., 201 McLean Blvd., Paterson, NJ 07504; (973) 881-8800, FAX: (973) 881-8361, e-mail: sales@synergymwave.com, Internet: www.synergymwave.com.** MRF

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1. Ulrich L. Rohde and D.P. Newkirk, *RF/Microwave Circuit Design for Wireless Applications*, Wiley, New York, 2000.
2. A.K. Poddar, S.K. Koul, and B. Bhat, "Millimeter Wave Evanescent Mode Gunn Diode Oscillator in Suspended Stripline Configuration," 22nd International Conference on Infrared and Millimeter Waves, pp. 265-266, July 1997.

Fast Synthesizers Switch 5 MHz To 20.48 GHz

These versatile, broadband frequency synthesizers employ a modular design to provide just those levels of performance needed in terms of step size, physical size, and frequency range.

Speed and silence are the trademarks of a high-quality microwave frequency synthesizer, and two of the outstanding attributes of the Series DS frequency synthesizers from Herley Communications Techniques, Inc. (Whippany, NJ). The single-enclosure synthesizers, which can be specified for coverage across any part of 0.005 to 20.48 GHz, feature typical switching speeds of 0.3 to 1 μs for step sizes as small as 1 Hz as typically as fast as 300 ns for step sizes from 250 kHz to 20 MHz. The phase noise is a very respectable—120 dBc/Hz at an offset of 10 kHz from a 10-GHz carrier.

ADRIAN LALICATA
Vice-President of Engineering
Herley Communications Techniques, Inc., 9 Whippany Rd., Whippany, NJ 07981; (973) 884-2580, FAX: (973) 887-6245, e-mail: alalicata@cti-inc.com, Internet: www.cti-inc.com.

The Series DS frequency synthesizers provide the response times and frequency agility required for a wide range of applications, including antenna testing, in satellite communications systems, for digital microwave radios, and in high-throughput automatic-test-equipment (ATE) systems. The modular design can accommodate almost any combination of frequencies within the total range, with resolution as fine or as coarse as required. Even with steps as fine as 1 Hz, the output frequency can be controlled to a precision of $\pm 1 \times 10^{-10}$.

Within the frequency range from 5.12 to 10.24 GHz, the Series DS synthesizers can be equipped with frequency step sizes of 1 Hz, 100 Hz, 250 kHz, 2.5 MHz, and 20 MHz. For lower-frequency coverage, the synthesizers can maintain the same step sizes or, as an option, can provide a reduction in step size by a factor of 2^N in relation to the output frequency. Standard models include the DS-120 with coverage from 2.56 to 10.24 GHz in 250-kHz steps, the DS-121 with coverage from 2.56 to 10.24 GHz in 1-Hz steps, the DS-161 with coverage from 640 MHz to 10.24 GHz in 1-Hz steps, and the DS-151 with coverage from 5 MHz to 10.24 GHz in 1-Hz steps.

The Series DS synthesizers produce +13 dBm output power. Spurious content is –80 dBc from 320 to 1280 MHz and –60 from 10.2 to 20.4 GHz. The single-sideband (SSB) phase noise is –92 dBc/Hz offset 100 Hz from a 10-GHz carrier, dropping to –120 dBc/Hz offset 100 kHz from the carrier and –128 dBc/Hz offset 1 MHz from the carrier. For a 1-GHz carrier, the phase noise is –111 dBc/Hz offset 100 Hz from the carrier and –147 dBc/Hz offset 1 MHz from the carrier.

Frequency can be programmed remotely via parallel TTL binary-coded-decimal (BCD) control signals. The synthesizers are available with a front-panel touch-screen display. **Herley Communications Techniques, Inc., 9 Whippany Rd., Whippany, NJ 07981; (973) 884-2580, FAX: (973) 887-6245, e-mail: alalicata@cti-inc.com, Internet: www.cti-inc.com.**

Comparator Measures Frequency And Phase

This precision frequency/phase comparator now incorporates a high-speed time-interval counter with generous memory and battery backup for demanding measurements.

JACK BROWNE
Publisher/Editor

Measurement of fractional differences in frequency and phase is often useful for evaluating sets of oscillators, precise lengths of cables, and other sets of high-frequency components. The model A7-X is an improved version of a proven frequency/phase comparator (the A7) from Quartzlock, with internal time-interval-counter card for handling measurements on devices with frequencies other than 5 or 10 MHz. The frequency/phase comparator can be used for testing minute differences in phase and frequency between frequency multipliers, dividers, standards, and other components.

Basically, the A7-X **(see figure)** employs a moving coil meter for rapid, unambiguous display of fractional frequency difference or relative phase difference between two sources. The comparator also provides outputs for connection to an external time-interval counter for high-resolution analysis of a source's time-domain stability (to 100 ps).

The major improvement in the A7-X is the inclusion of an internal model A7-X TIC time-interval-counter card, which eliminates the need for an external counter. Supported by Windows-based software, the counter's resolution is around 50 fs for a single measurement with filter off, and around 15 fs with 10-Hz filter.

The counter acquires data at a rate of 1000 readings/s, with data analysis performed by the Stable 32 PC software. Data can be saved in the form of text files for further analysis. The rapid data-acquisition rate supports phase-noise calculations to a Fourier frequency of 500 Hz from the carrier. By extending the data acquisition to several thousand seconds, phase noise can be analyzed within 1 mHz of the carrier.

The frequency/phase comparator has two modes of operation: frequency-measurement mode and phase-difference mode. In the first, the moving coil meter indicates fractional frequency difference and the counter operates like a frequency counter, with full-scale meter ranges from $\pm 10^{-7}$ to $\pm 10^{-12}$. The root-mean-square (RMS) resolution is better than 5×10^{14} for a 1-s gate. In the second mode, the meter reads the relative phase difference between the reference and the measurement inputs, with a full-scale range selectable between ±10 µs and ±10 ps.

Two versions of the A7-X frequency/phase comparator are currently available: the analog model A7-A offers 1×10^{-13} measurement resolution while the metrology grade model A7-MX) provides resolution to 1×10^{-16}. **Quartzlock UK Ltd., Gothic, Plymouth Rd., Totnes, Devon TQ9 5LH, England; (44)(0)1803 862 062, FAX: (44)(0)1803 867 962, e-mail: quartzlock@quartzlock.com, Internet: www.quartzlock.com.**

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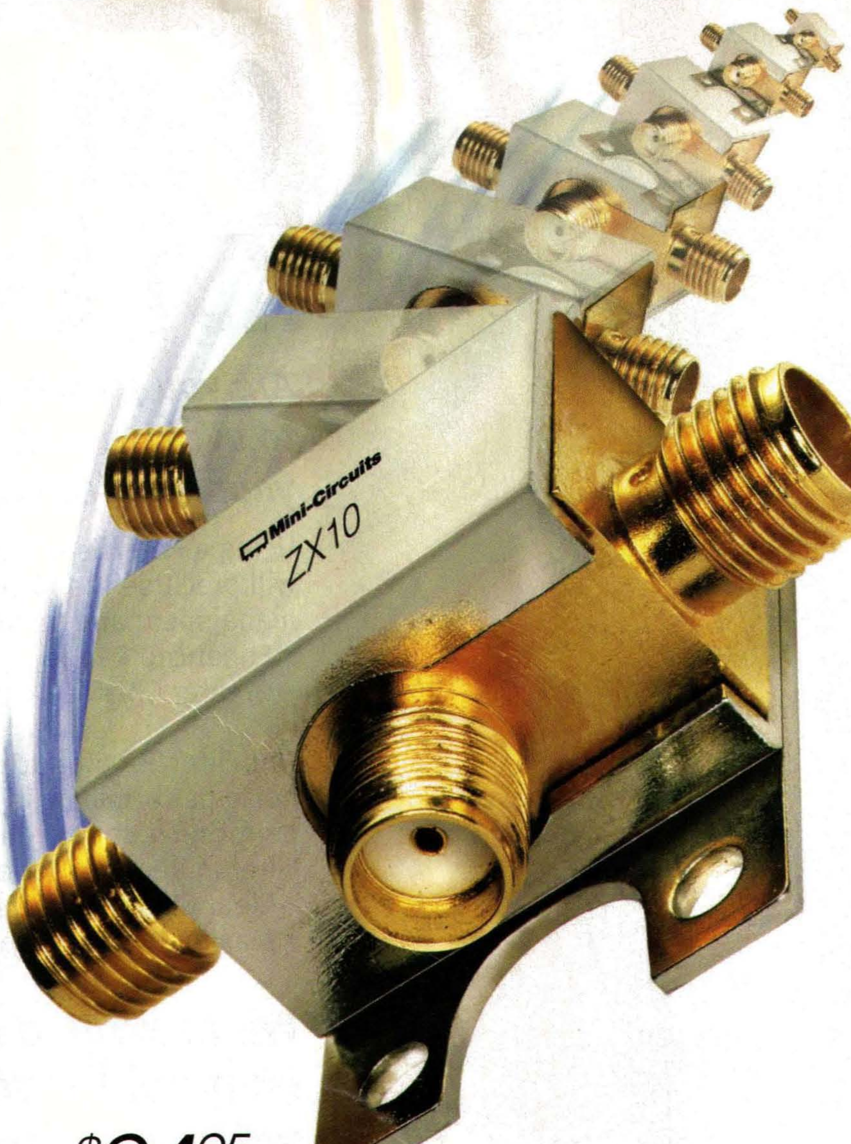

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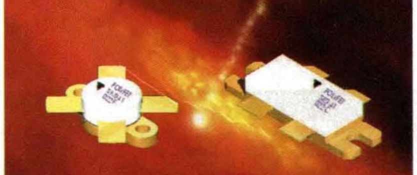

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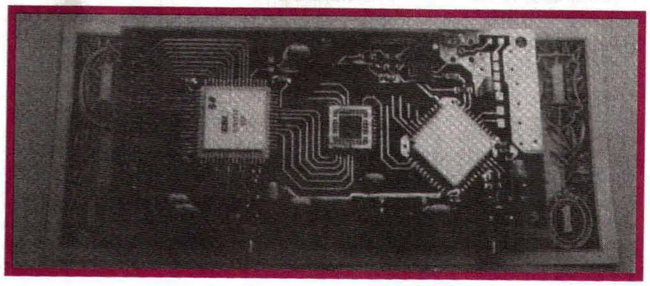
looking back



THIS JULY MARKS THE 16th anniversary of a breakthrough product from Sciteq Electronics (San Diego, CA), a company then synonymous with direct-digital-synthesizer (DDS) technology. The company's ADS-2 synthesizer tuned from 0.1 to 250 MHz with 20-ns frequency switching speed and operated at clock rates beyond 1 GHz.

next month

Microwaves & RF July Editorial Preview

Issue Theme: Amplifiers & Oscillators

News

July's issue beats back the boredom of hot summer days with a Special Report on power combiners/dividers. Although often taken for granted as a passive component by higher-order designers, power combiners and designers are essential building blocks for modern high-power amplifier (HPA) designers. This exclusive survey will provide the basics of design, examine the current state of the art along with recent technological advances, and offer a comprehensive listing of RF/microwave power combiner/divider manufacturers along with a sampling of their wares..

Design Features

July begins one of the most ambitious article series in magazine history: an eight-part design-feature collection on amplifier design. Authored by Joe White of JFW Technology (and founder of *Applied Microwave* magazine), this series *is guaranteed* to teach readers how to design microwave amplifiers, starting with the basics of using S-parameters in Part 1. Additional technical articles in July include a report on the next generation of logarithmic amplifiers for wide-dynamic-range commercial and military receivers, a method for generating ultrawideband (UWB) arbitrary waveforms by means of optical spectrum sculpting, how to design a tracking generator for a microwave spectrum analyzer. Also, an author from Infineon details an effective bias circuit for LDMOS power transistors.

Product Technology

July ushers in the latest version of a powerful electronic-design-automation (EDA) software suite tailored to the RF/microwave designer. The latest upgrade features new active and passive models, more powerful solvers, and a foolproof user interface. Additional product features in July offer some of the highlights of the recent MTT-S, including a self-contained VCO test set that can measure phase noise on carrier signals from 10 MHz to 7 GHz at levels approaching the thermal noise floor, a handheld spectrum analyzer with 6-GHz range, and a broadband power amplifier that delivers a hefty 10 W output power from 2 to 20 GHz.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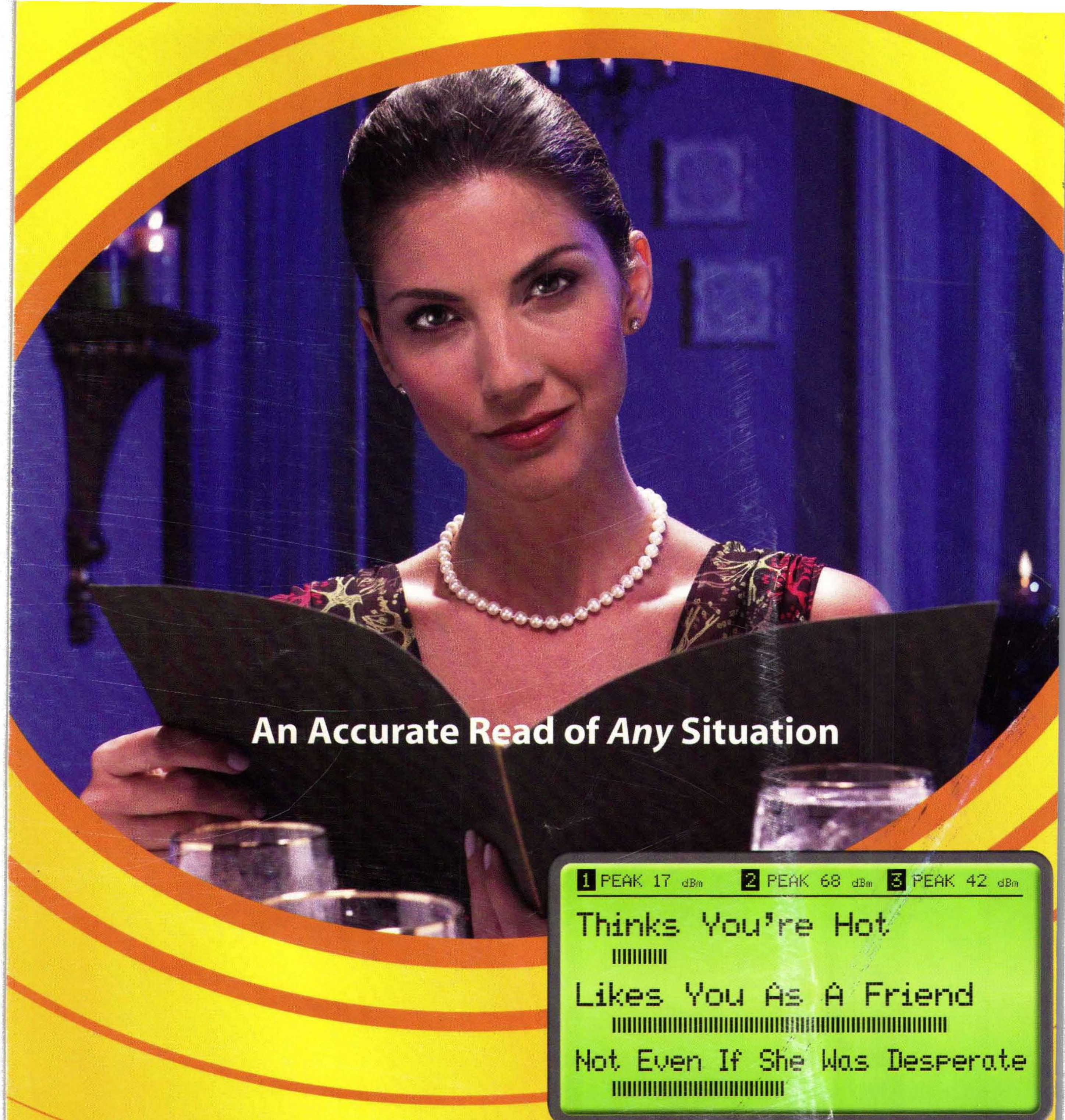
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